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EXPERIMENTAL TRANSMITTING AND RECEIVING EQUIPMENT FOR HIGH-SPEED FACSIMILE TRANSMISSION

V. SYNCHRONIZATION OF TRANSMITTER AND RECEIVER

by D. KLEIS and M. van TOL.

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The motors which bring about the scanning of the picture areas in the transmitter and in the receiver of the Philips apparatus for high-speed facsimile transmission have to run synchronously with a tolerance of no more than 0.6 degree in their relative phase. This requirement cannot be met with synchronous motors, nor with the system of stabilization by means of tuning-fork generators much used for low-speed facsimile transmission. A new method of synchronization had, therefore, to be developed. A regulating device was employed which reacts to phase deviations between the motors in the transmitter and the receiver, combined with a similar device reacting to differences between the speeds of the two motors. These devices are controlled by synchronizing pulses produced by the optical rotor in the transmitter and transmitted to the receiver together with the facsimile signal, and with the aid of pulses produced by a small generator coupled to the shaft of the receiver motor. Moreover, in the transmitter a device is used which stabilizes the speed of the transmitter motor, so that the synchronization only needs to provide a correction for small variations in the transmitter. With this method it has been possible to reach the necessary phase constancy; in the event of a disturbance the equilibrium is aperiodically restored, with such an inertia that the edge of the recorded picture does not show any disturbing undulation.

Synchronization is a problem of prime importance in all systems of picture transmission, both in facsimile telegraphy and in television. On the transmitting side there is a picture surface traversed by a scanning spot, and at the receiving end a picture surface traversed by a recording spot. Both surfaces are identical and one can imagine them as being divided equally into minute surface elements. To reproduce the transmitted picture in the receiver without any geometrical distortion, the scanning spot and the recording spot must be "synchronized", which means that they must move continuously over corresponding areas of the two picture surfaces at exactly the same moment [1]).

The picture surface is always scanned along parallel, adjacent lines. We can therefore express the synchronization condition in a somewhat more concrete form by saying that the scanning spot and the recording spot have to start traversing each line simultaneously and must take the same length of time to cover a line. If the recording spot starts sometimes too early and sometimes too late then the lines in the picture obtained show individual displacements in their own direction; see *fig. 1.* It is not necessary (and not possible) to exclude such displacements entirely, but the tolerances in this respect are rather small; with a picture width of say 20 cm a maximum line displacement of 1 mm, i.e. 1/200 of the line length, is admissible. Of course such a displacement cannot be allowed between adjacent lines; there must be a gradual transition, so that the edge of the picture shows an undulation with a not too small wavelength. As a normal requirement it is taken that this "wave" must have a wavelength of at least 4 cm (fig. 1).

For the numerous systems of facsimile telegraphy already existing a number of methods of synchronization have been worked out, none of which however was suitable for our new facsimile system working at a transmission speed about 60 times higher than the existing systems. Some simple quantitative considerations will make this clear. The method of synchronization devised for this new system will then be described on broad lines.

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[1]) The transmitted signal has always a finite though very short transit time, *T*. On closer examination it is seen therefore that the recording spot, the intensity of which is governed by the signal obtained from the scanning spot, must traverse the picture elements with a delay *T* with respect to the scanning spot. It is not therefore absolutely correct to say that the scanning and recording spots pass over corresponding elements of the two picture areas at exactly the same moment, but this is of no consequence whatever for our further considerations.

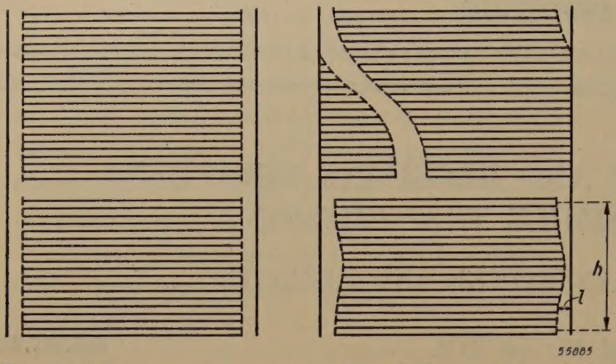

Fig. 1. Diagrammatic representation of the position of the parallel lines scanned on the picture area in the transmitter (left) and recorded on the picture area in the receiver (right). The fully-drawn and the broken lines at the edge indicate the picture area and the actual picture respectively. In the time required to traverse the short distance between these two lines the transmitter does not emit any picture signal. *Upper half:* An example of inadequate synchronization. Here the receiver has a phase disturbance which for some length of time increases and then remains constant. *Bottom half:* The synchronization is considered adequate if the displacement of the picture edge (variation of the distance l) does not exceed 1/200 line length, i.e. about 1 mm, and the undulation of the edge has a wavelength (h) of at least 4 cm.

For the details of the construction of the transmitter and the receiver in the Philips high-speed facsimile system we refer to the four previous articles published in this journal, which will be referred to as I-IV [2]).

Synchronous motor and tuning-fork oscillator

Since the synchronization can be reduced to the producing of an identical frequency at the transmitter and receiver, it seems obvious to take advantage of the standard frequency at our disposal in the a.c. mains. In that case the movements of the scanning spot and the recording spot are both brought about by synchronous motors fed from the mains. This method is of course confined to those cases where the mains available for the transmitter and for the receiver are interlinked and therefore always have the same frequency. In such cases this method is indeed successfully applied for low-speed facsimile systems.

Why is this not possible with our system?

In low-speed facsimile transmission the scanning rate is usually about 3 lines per second. The driving two-pole synchronous motors (the most favourable for our purpose) run at the rate of 50 r.p.s. (3000 r.p.m.) in Europe and 60 r.p.s. (3600 r.p.m.) in America, thus making about 20 revolutions in scanning one line. For a permissible displacement of 1/200 line length between transmitter and receiver a phase displacement of $20 \times 360°/200 = 36°$ can be allowed between the motors at the transmitting and receiving ends. The phase displacements between the terminal voltages of the two local mains likely to occur through fluctuations in the instantaneous load will as a rule be well below this limit of 36°. If, therefore, the mutual phase of the scanning elements in transmitter and receiver is properly adjusted at the beginning of the transmission, no prohibitive displacements need be feared.

With our high-speed facsimile system the rate of scanning is 180 lines per second (see article II). A driving synchronous motor would therefore make only $^1/_3$ revolution in scanning one line. The permissible phase displacement between the motors at the transmitting and receiving ends would thus not be more than $^1/_3 \times 360°/200 = 0.6°$. One could not possibly reckon on such a phase constancy of the mains.

Consequently synchronous motors cannot be considered for our purpose, not even if it could be expected that interlinked mains would be available at the transmitting and receiving ends.

In very many cases, especially in international communications, this latter condition will certainly not be complied with and for this reason other methods of synchronization have had to be developed already for low-speed facsimile systems. Practically all these methods result in the generation of an oscillation with a frequency of great constancy at the transmitter and receiver, this oscillation being used to control the speed of the motor for the scanning device. The two frequencies are rendered as accurately identical as possible by local adjustment of the frequency-governing element, for instance a tuning fork.

Whereas with the method of synchronous motors the mean speeds are identical and only disturbances due to phase shift fluctuations have to be considered, with these other methods the possible small frequency differences of the local tuning fork oscillator must be considered. The smallest difference in frequency will result in an additional small phase shift of the motors in each cycle. After a large number of cycles this displacement will have accumulated considerably. Suppose that the tuning forks vibrate at such a frequency that p cycles occur during the scanning of one line. A picture to be transmitted may be 30 cm long, so that given a line width of 0.2 mm the picture may contain in all 1500 lines. Each tuning fork then makes 1500 p vibrations per picture. If at the end of the picture the line displacement is not to exceed the value of 1/200 line length, the number of vibrations of the tuning fork must not be more than $p/200$ out in the 1500 p, that is to say the frequency deviation must not be more than 1 : 300,000.

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[2]) Experimental transmitting and receiving equipment for high-speed facsimile transmission.
- I, General, by H. Rinia, D. Kleis and M. van Tol, Philips Techn. Rev. **10**, 189-220, 1948 (No. 7).
- II. Details of the transmitter, by D. Kleis, F. C. W. Slooff and J. M. Unk, Philips Techn. Rev. **10**, 257-264, 1948 (No. 9).
- III. Details of the receiver, by F. C. W. Slooff, M. van Tol and J. M. Unk, Philips Techn. Rev. **10**, 265-272, 1948 (No. 9).
- IV. Transmission of the signals, by D. Kleis and M. van Tol, Philips Techn. Rev. **10**, 289-298, 1948 (No. 10).

This requirement can reasonably be met witb tuning forks or other vibrating elements if they are very carefully made, but this will not bring us any farther than ensuring that one picture is properly recorded. In order to avoid the consecutive pictures being displaced farther and farther across the picture area, the phase of the scanning device in the receiver has to be corrected after every picture. With the usual low-speed facsimile transmission there is no difficulty in having this correction made for instance by the operators, but in our case, when it takes no more than 8 seconds to transmit a picture of the length mentioned, the operators would have to make this correction every 8 seconds, which of course is out of the question. Or, to put it the other way round, to allow of a correction being made say only once every 10 minutes, a frequency constancy of 1 in 2×10^7 would be required. Even with the best means available (quartz oscillators in thermostats) this cannot be attained.

There are two possible ways of overcoming these difficulties. According to the first method one is satisfied with a frequency constancy merely sufficient, as described above, for one picture or even less, and an additional mechanism is provided which after every picture or at the required shorter intervals automatically controls the phase of the scanning device in the receiver and where necessary makes a correction to annihilate the phase deviation. By the second method speed is not governed by local tuning devices — which would not possess sufficient frequency constancy anyhow — but the speed of the receiver motor is controlled by a frequency which is derived from the transmitter motor and transmitted together with the picture signal.

The synchronization in our facsimile system may be regarded as a combination of these two method s It will be seen why neither the first nor the second method alone will suffice.

Stabilizing the speed of the transmitter motor

If one has the means of getting the speed of the receiver motor to follow exactly any speed variations of the transmitter motor then, in principle, it is not essential that the transmitting speed should be kept to a high degree of constancy. However, the smaller the speed variations of the transmitter, the easier it will be, of course, to follow them at the receiver with the necessary accuracy. For this reason, in our facsimile equipment steps have been taken to stabilize the speed of the transmitter motor, and it is well to describe this stabilization first.

The transmitter motor is a direct current motor, the armature of which is fed from a simple (non-stabilized) mains rectifier, whilst the field current is supplied by an output valve of the type EL6. The speed of such a motor is approximately inversely proportional to the magnetic flux in the armature. The motor can therefore be accelerated by reducing the field current and slowed down by increasing the field current. For the normal speed of 60 r.p.s. the current to be supplied by the output valve has to be 55 mA, for which the grid voltage of the valve has to be about —8 V. To stabilize the speed it is arranged that the grid potential of the output valve becomes more positive as the motor begins to rotate at a higher speed. The resulting increase of the anode current flowing through the field winding then counteracts the increase in speed.

This regulating voltage at the grid of the output valve is derived in the following way. On the shaft of the motor a small generator is fitted which at a speed of 60 r.p.s. produces an alternating voltage with a frequency of (for instance) 180 c/s. This alternating voltage is applied to an L-C circuit tuned to a frequency slightly higher than 180 c/s, such that the resonance curve has its steepest slope at 180 c/s; see *fig. 2*. Consequently the amplitude of the voltage across the circuit will vary considerably when the frequency of the alternating voltage supplied, that is to say the speed of the motor, changes but slightly. The voltage across the circuit is now rectified, and the direct voltage obtained — from which is to be subtracted a constant bias determining the working point — serves as control grid voltage of the output valve. A block diagram of this stabilizing circuit is given in *fig. 3*.

Let us now consider the working of this stabilization more closely. In the event of an interference, for instance a slight variation in the armature voltage or in the load of the motor, which has the tendency to change the speed, the field current changes to such an extent as to compensate the effect of the interference very rapidly. This means that the equilibrium corresponding, say, to a changed armature voltage, is already obtained with an extremely small change in speed. An actual and permanent small change in the speed, however, is indispensable. The effect of the stabilization therefore is only that the speed change required for the new equilibrium is considerably reduced as compared with the case where there is no stabilization.

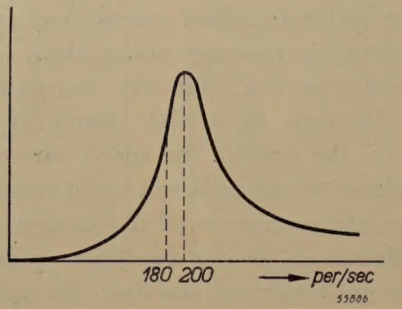

Fig. 2. Resonance curve of the tuned circuit employed for stabilizing the speed of the transmitter motor.

In our installation the reduction factor has been made very large by keeping the damping of the *L-C* circuit low, namely by limiting the damping effect of the coupled rectifier with the aid of a special circuit (cf. fig. 6). In this way a resonance curve is obtained with steep sides, and the anode current of the output valve (i.e. the field current of the motor) will therefore greatly vary with the speed, due also in part to the high mutual conductance of this valve. In our case, for an interference due to variation of the armature voltage, the reduction factor is 270. For a variation of the load it is still higher.

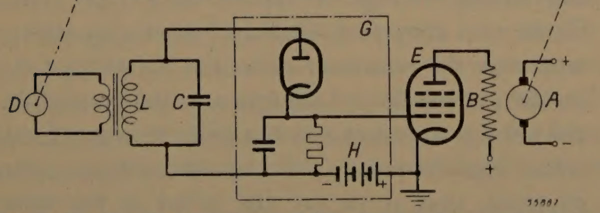

Fig. 3. Circuit diagram for stabilization of the speed of the transmitter motor. *A* armature fed from a mains rectifier; *B* field winding excited by the anode current from the output pentode *E*. On the motor shaft is a small generator *D* producing an alternating voltage which causes the tuned *L-C* circuit to oscillate. The voltage across the circuit is rectified by the rectifying circuit *G* and the direct voltage thus obtained, which varies considerably with the speed of the motor, is applied to the grid of the output valve. As the speed increases, the anode current of *E* rises and the stronger excitation slows down the armature. The right working point on the characteristic of the output valve is adjusted with the grid bias *H*.

The effect of a change in the natural frequency of the *L-C* circuit, on the other hand, cannot be reduced by this stabilization, as will be readily seen. It can therefore be said that the degree of constancy of the speed is practically determined by the constancy of the values of L and C in the circuit. For our purpose a variation in the speed over long periods to the extent of 1% or more could be allowed, so that no very stringent requirements had to be made as far as this circuit is concerned.

Other points of importance in the stabilization, such as automatic starting, inertia of adjustment, the possibility of instability arising, etc., cannot be dealt with here. It is to be noted, however, that the nominal speed can be altered by changing the capacitor in the *L-C* circuit. Another means by which small changes could be made in the nominal speed would be to change the abovementioned constant negative grid bias of the output valve, which is added to the regulating voltage. We shall presently see why this is of importance.

Regulating the speed of the receiver motor

In the first place the receiver is equipped with a speed-stabilizing device very much resembling that in the transmitter. The motor is of the same type; the field current is supplied by an output valve; the latter is controlled by the rectified voltage of a tuned circuit which in turn is excited by an alternating voltage of 180 c/s obtained from a small generator on the motor shaft. Here again negative grid bias is added to the regulating voltage for correctly adjusting the working point of the valve.

The result of this stabilization is that the effect of mains voltage fluctuations, etc., on the speed of the receiver motor is very greatly reduced, as is the case with the transmitter. At the same time, however, it affords a very simple means of satisfying the desire for the *speed of the receiver to be continuously controlled by that of the transmitter*. For this purpose an alternating voltage of a frequency proportional to the speed of the transmitter motor is transmitted by cable from the transmitter to the receiver [3]. There this voltage is applied to a tuned circuit corresponding to that in the transmitter, and the rectified voltage of this circuit is used for stabilizing the receiver motor, instead of the constant grid bias of the output valve. At the end of the previous section we have already pointed out that with the regulating method described the nominal speed can be adjusted by choosing a suitable value for this grid bias. Now in this manner a variation in the speed of the transmitter will cause also a variation of the nominal speed at which the stabilization of the receiver is working. A block diagram of this method of synchronization is shown in *fig. 4.*

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[3] One might use for this the already mentioned voltage of the generator on the shaft of the transmitting motor. Actually we use the "synchronization signals" referred to already in articles II and IV. This point is considered further in what follows.

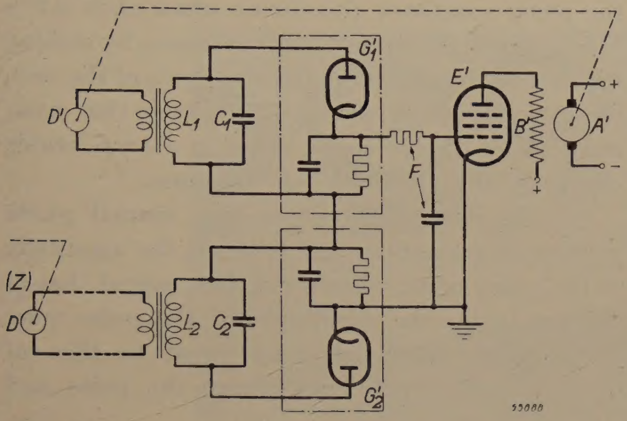

Fig. 4. Block diagram of the speed regulation of the receiver motor. The system is based upon a similar method of stabilization of the speed as applied in the transmitter (fig. 3): the small generator D' is mounted on the shaft of the armature A' of the receiver motor; the alternating voltage generated acts upon the circuit L_1—C_1; the circuit voltage rectified by G_1' controls the output valve E', the anode current from which flows through the field winding B'. Instead of a constant grid bias (H in fig. 3) for the output valve, here the voltage from the circuit L_2—C_2 rectified by G_2' is used, this circuit being caused to oscillate by means of an alternating voltage (Z) derived from the rotation of the transmitter motor and transmitted together with the picture signal. F is a smoothing filter.

However, our object in view is still not by any means reached, as the reader may already have observed. The receiver now reacts to a small change in speed of the transmitter, but in the event of an interference in the receiver itself, for instance a mains fluctuation, there will be an independent change in speed which is necessary for the new equilibrium and which, in spite of the large reduction factor of the stabilization, may be much greater than is permissible. Moreover, the influence of the local frequency-governing elements, which in the foregoing led to the rejection of the method of synchronization with tuning-fork generators, is not by any means eliminated. In fact there is no actual synchronization here; the control is not affected by the question whether the speeds in the transmitter and the receiver are actually equal; it only causes the speed of the receiver to vary in the corresponding degree to that of the transmitter. The nominal speeds of transmitter and receiver are governed by the local L-C circuits, which by no means possess the necessary frequency constancy of 1 in 2×10^7, so that after a short time, even in the absence of other disturbances, prohibitive phase displacements would arise.

The solution has been found in the principle of the other means already outlined: an automatic control, in our case functioning continuously, which has a direct phase-correcting action. This we shall now proceed to describe. We shall leave for the moment the question what function is then performed by the speed control already dealt with.

Regulating the phase of the receiver motor

As already stated, the receiver motor drives a small generator supplying the alternating voltage required for stabilizing the speed. This generator consists of a fixed magnet with an inductor and a rotating armature with three teeth. The teeth are so shaped that as one of them passes the magnet a voltage pulse is generated in the inductor. When the motor is running at the nominal speed of 60 r.p.s. there are thus 180 pulses per second. The fundamental frequency of 180 c/s filtered out of these pulses is used for the speed-stabilizing system (fig. 4). The impulses themselves are required for the phase control now to be described.

The synchronizing alternating voltage of the transmitter, mentioned above, also consists of pulses. These are obtained by optical means, i.e. by the three rotating, optical scanning systems in the transmitter (see article II): alongside the edge of the document to be scanned is a somewhat specular aluminium plate; an pulse of constant height and duration is thus introduced in the facsimile signal at the beginning of each scanning line. These pulses, which serve not only for the synchronization but also for reconstructing the scale of blackness of the picture (see article IV), are transmitted along the cable together with the signal; thus no separate channel is needed for the synchronization signals.

To achieve the speed regulation in the receiver already described (fig. 4) the fundamental frequency of 180 c/s is again filtered out of the transmitted pulses and used. For the phase regulation, however, the transmitter pulses are converted by means of a simple network into a saw-tooth voltage, and to this are added the pulses of the receiver generator. When the two motors are running at the same speed a purely periodical voltage is obtained in the form as represented in *fig. 5.*

The position of the pulse on the "back" of the saw-tooth is apparently directly related to the mutual phase of the transmitter and the receiver. By suitably positioning the inductor in relation to the armature on the shaft of the receiver motor it can be arranged that for the desired phase relation (the recorded line in the receiver beginning at the edge of the picture plane) the pulse just comes to lie midway between the fly-back surges of the saw-teeth.

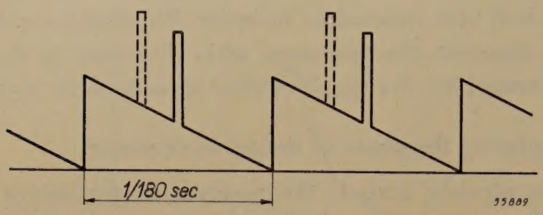

Fig. 5. Saw-tooth and periodical pulse added together. The saw-tooth voltage is derived from the rotation of the transmitter motor, the pulse voltage from the rotation of the receiver motor. When the motors are running at the same speed the pulse takes up a fixed position on the "back" of the saw-tooth; this position depends upon the phase relation of the motors. When the phase relation changes, the pulse is displaced along the saw-tooth, its top becoming higher or lower (for instance as shown by the dotted lines).

This voltage according to fig. 5 is conducted to a rectifying circuit which supplies a voltage equal to the difference between the peak of the pulse and the mean value. This direct voltage is applied, in series with an adjustable bias, to the grid of an output valve the anode current of which forms part of the field current for the receiver motor; see *fig. 6*.

In the event of the receiving motor lagging in phase behind the transmitting motor for some reason or other, the pulses come to lie farther to the right on the saw-tooth. Thus, its top lies lower, the output voltage of the rectifier is smaller, and when the polarity is suitably chosen the grid voltage of the output valve becomes more negative and as a consequence the field current of the motor is reduced. The result is that the motor is accelerated, so that it begins to make up for the phase lag in relation to the transmitter motor. In this way the original phase relation between transmitter and receiver is restored.

The equilibrium is also determined in part by the grid bias of the regulating valve. By varying this bias the phase relation between transmitter and receiver can be adjusted at the beginning of a transmission. For this purpose the potentiometer with which the grid bias is varied is provided with a control knob on the front panel. Also mounted on the front panel is a meter indicating the phase difference, whilst furthermore the form of the voltage illustrated in fig. 5 can be checked with the aid of a small cathode-ray oscilloscope. These devices are shown in the photograph reproduced in *fig. 7*.

If the cause of a phase displacement persists then it is not possible for the phase to be fully restored, just as we have seen with the methods of regulation first described. The effect of an interference is only reduced, that is to say the new equilibrium is already obtained at a very small displacement of phase. It is found possible to make the reduction factor sufficiently large to satisfy our demands as regards the phase constancy (maximum phase shift 0.6°). This is achieved, apart from other means, by making the saw-tooth steep, thus the amplitude of the saw-tooth voltage high (about 250 V), so that even small phase displacements result in a considerable change in the amplitude of the pulses.

For the control to act at any mutual phase position of transmitter and receiver, the amplitude of the pulse must, wherever it lies, exceed that of the saw-tooth; the amplitude of the pulse must therefore be made at least equal to that of the saw-tooth. The amplitude of the pulse and

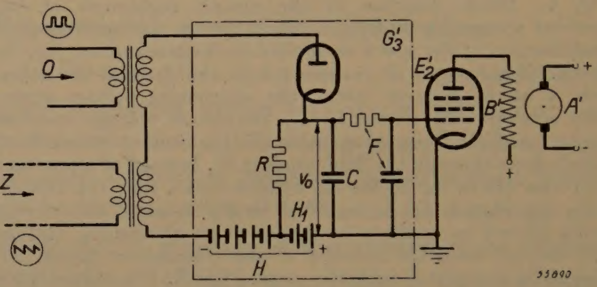

Fig. 6. Diagram of the phase regulation. Again an output valve is used (E_2') to supply a part of the current for the field winding B' of the receiver motor. This valve is controlled by the output voltage of the rectifier G_3', this being proportional to the difference between the peak value and the mean value of the A.C. input voltage. The input voltage for the rectifier is the voltage illustrated in fig. 5; part of this voltage is supplied by the receiver (O) and part of it by the transmitter (Z). The variable bias H ensures that the voltage v_0 in the right phase position assumes the value of -8 V corresponding to the desired working point of the output valve.

The rectifying circuit differs somewhat from the simple case illustrated in figs 3 and 4. It has the advantage that the voltage across the resistor R is reduced to the difference between the voltage v_0 and the part H_1 of the base, i.e. in the undisturbed state about 7 V, as against v_0-H, i.e. for instance 370 V, with the normal system. Hence, of the total variation range (250 V) of the peak height of the impulses — independently of their undisturbed position on the back of the saw-tooth — for the regulating effect only a part amounting to H_1 (= 15 V) is used, which just matches the grid base of the output valve. In this case too, the ripple of the grid voltage is only proportional to the voltage across R instead of the peak value of the inpulses. Consequently the RC-constant of the rectifying system (and also of the smoothing filter F) could be made sufficiently small to render the delay of the regulation described below innocuous.

Thanks to the reduction of the voltage across R much less energy is dissipated in this resistor than is the case with the normal system. It is for this reason that the principle of the system sketched is actually also applied for the rectifiers in figs 3 and 4. The smaller dissipation is of importance because it is accompanied by less damping and thus gives a sharper resonance curve for the preceding L-C circuit, as has already been pointed out above.

that of the saw-tooth voltage must be kept accurately constant, because any variation would just as well cause the voltage across the rectifier to change as would a phase displacement. Consequently the incoming pulses of the transmitter and the receiver are not applied directly but first limited to a constant level in the usual way.

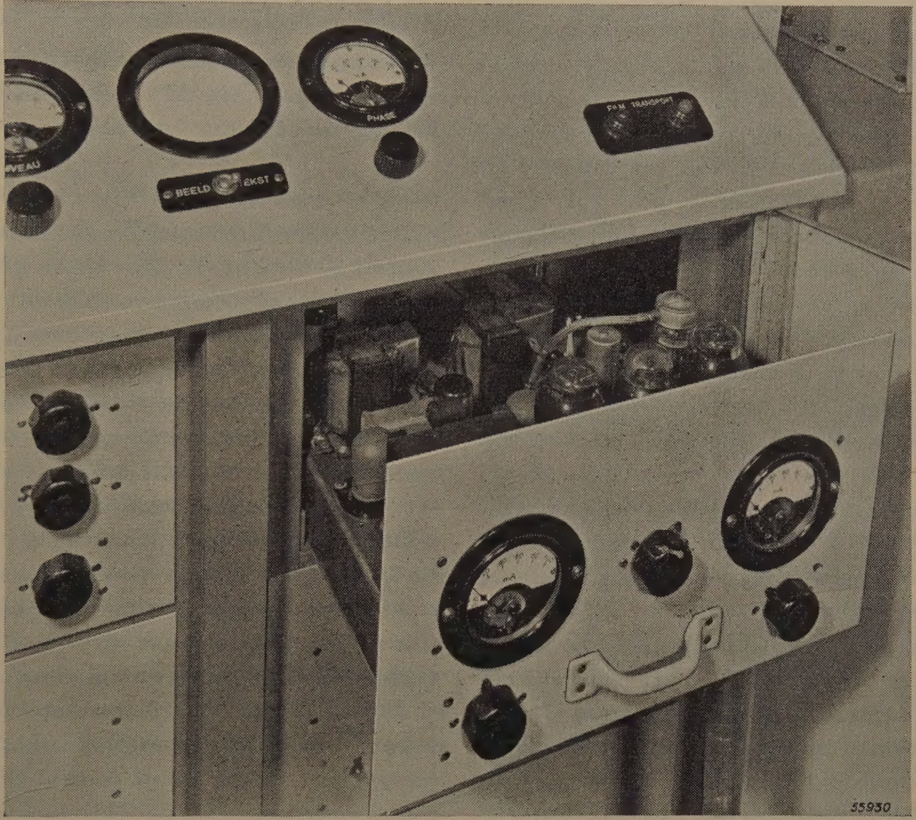

Fig. 7. The synchronization system for the receiver is for the greater part mounted on a single chassis which slides into the receiving cabinet. On the slanting top panel of the cabinet can be seen the meter and the control knob for reading and adjusting the phase relation between transmitter and receiver. To the left of these is the screen of the cathode-ray oscilloscope on which the oscillogram of the voltage for the phase regulation (fig. 5) can be produced by closing the switch. On the front panel of the synchronization chassis one sees the potentiometer knob and the two milliammeters with which the contributions from the two output valves of the speed and phase regulations towards the field current are adjusted (see caption of fig. 8). Immediately behind the front panel are the output valves, between which is one of the amplifiying valves for limiting the pulses supplied by the transmitter and the receiver.

Necessity of regulating speed in addition to phase

It is readily understood that phase regulation automatically implies that the speeds of the transmitting and receiving motors must be strictly equal. The smallest difference in speed manifests itself as a continuously increasing phase displacement. Owing to the pulse being shifted farther and farther on the saw-tooth this phase displacement in turn gives rise to an increasing deviation of the field current, and with a constant speed difference the correcting torque on the motor therefore becomes larger and larger. Thus the speed difference is neutralized and its cause can only bring about a displacement of the equilibrium to a somewhat different phase relation, as we have already seen above [4]).

It may now be asked why, then, this phase control alone does not suffice.

Phase control can only be effective when the speeds of the transmitter and receiver are already practically equal. If there is any appreciable difference in the speeds then the pulse runs very rapidly over the back of the saw-tooth right to the end and then begins again at the other end. The rectified voltage would therefore have to rise and decay at the same rate. Owing to the necessary smoothing of the output voltage of the rectifier, however, the grid voltage of the output valve corresponds to the average peak height, and consequently the total effect of the rapid phase variations upon the speed of the motor is nil. (As a matter of fact this would also be the case without smoothing, because the inertia of the armature would render rapid variations of the torque ineffective.)

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[4]) In the terminology of control technique the phase control might be termed an "integral control" of the speed. Integral controls have the property of being able to make the mean value of the quantity to be adjusted exactly equal to a prescribed value.

The incapability of the phase control to correct large speed deviations is particularly manifest in the fact that when the receiver motor is started it will as a rule not be brought up to the desired speed. This is actually done by the speed control. After being switched on the receiver motor starts running and tries to reach the rated speed to which the speed stabilization is set. The speed regulation by means of the synchronization signals corrects this in so far that it brings the speed of the receiver anyhow very close to the speed of the transmitter. This means that the pulse in fig. 5 runs more and more slowly over the saw-tooth until the phase control comes into operation and keeps the pulse in a certain position. The phase relation can then be further adjusted by hand, as explained above.

Though easy starting is in itself sufficient justification to apply speed control in addition to phase control, there is a second reason for this which we regard as being still more important. In order to explain this reason we must first look at one side of the problem that has so far been neglected, namely the behaviour of this regulation as a function of time. In the description given above we have spoken only of the state of equilibrium, but how is a new equilibrium attained after an interference has taken place?

If the angle of rotation of the transmitting motor with respect to an arbitrary initial position is Φ, then in the event of an interference the angle of rotation of the receiver motor with respect to the corresponding initial position will not be exactly equal to Φ, but $\varphi = \Phi + \psi$. Thus ψ is the undesired phase deviation. Now there are a number of torques acting upon the shaft of the receiver motor. The normal field present when there is no disturbance in the running of the motor supplies a driving torque M_0. Friction sets up a braking torque, which for our purpose we can express to a sufficient approximation by $M_r = a + b\,\dot{\varphi}$, in which a and b are constants. The phase control yields a correcting torque M_c, which is opposed to the phase displacement ψ and which for small deviations we can take as being proportional to ψ: $M_c = c\psi$. Similarly the speed control yields a correcting torque M_d proportional to the speed difference $\dot{\varphi} - \dot{\Phi}$, thus $M_d = d\dot{\psi}$.

Using I to denote the moment of inertia of the armature and the rotor of the scanning device coupled to it, for the rotation of the receiver motor we have the differential equation:

$$I\,\ddot{\varphi} - M_0 + (a + b\,\dot{\varphi}) + M_c + M_d = 0.$$

Substituting $\varphi + \Phi + \psi$ and putting $\ddot{\Phi} = 0$, since we shall assume the transmitter motor to rotate at a constant speed, we then have

$$I\,\ddot{\psi} - M_0 + a + b\dot{\Phi} + b\dot{\psi} + M_c + M_d = 0.$$

This equation must also hold for the undisturbed state where ψ, $\dot{\psi}$, $\ddot{\psi}$ M_c, M_d are all nil. Hence $M_0 = a + b\dot{\Phi}$, so that our equation becomes

$$I\,\ddot{\psi} + b\dot{\psi} + M_c + M_d = 0. \quad . \quad . \quad . \quad . \quad (1)$$

Let us now first consider the effect of the phase control alone. For the time being we therefore take the term M_d as being zero. Substituting $M_c = c\psi$, we then see that eq. (1) assumes the form of the well-known equation for a damped oscillatory vibration without an external force acting upon the vibrating system. From this we can at once conclude that when an arbitrary deviation ψ is introduced and the system is further left to itself the relative phase ultimately returns to its original value ($\psi = 0$). The relation of the damping term (coefficient b) to the other terms will determine whether this return of the relative phase to its original value takes place in an oscillatory manner or aperiodically.

This, however, is still incomplete. We have to take into account the fact that both in the phase control and in the speed control there is a delay time τ respectively τ_1. This is for a large part attributable to the RC-constant of the rectifying circuits supplying the grid voltage for the regulating valves and that of the smoothing filters, and also for a part to transients, for instance in the tuned circuits, and to the hysteresis of the iron circuit of the motor. The delay means that the correcting torques M_c and M_d at the instant t are determined by the phase and speed deviations respectively which were present at a time τ respectively τ_1 earlier. Thus:

$$M_c\,(t) = c\,\psi\,(t - \tau); \qquad M_d\,(t) = d\,\dot{\psi}\,(t - \tau_1).$$

If we now develop both these expressions in progressions according to Taylor we get:

$$M_c\,(t) = c\,[\psi\,(t) - \tau\,\dot{\psi}\,(t) + {}^1/_2\,\tau^2\,\ddot{\psi}\,(t) - \dots],$$
$$M_d\,(t) = d\,[\dot{\psi}\,(t) - \tau_1\,\ddot{\psi}\,(t) + \dots].$$

For these qualitative considerations we may ignore higher terms of the series. By substituting these expressions we derive from eq. (1):

$$(I + {}^1/_2\,c\tau^2 - d\tau_1)\,\ddot{\psi} + (b - c\tau + d)\,\dot{\psi} + c\psi = 0 \dots (2)$$

As is to be seen, the coefficients of the equation for the vibration are altered owing to the retardation effects. In particular it is to be noted that now the damping term may also become negative (b, c and d according to their definition are positive), namely when c is too large. The phase deviation ψ will then fluctuate around the zero value; but the fluctuations will become larger and larger. This phenomenon may indeed occur in practice and would, of course, render the control useless. This is all the more unwelcome because a high value of the coefficient c, indicating the "amplification" of the phase regulation, is desired for a large reduction factor in this regulation.

This now, is where advantage is again taken of the speed regulation. The contribution of this regulation towards the damping term consists in the positive quantity d appearing in eq. (2). By making d large, we can therefore always make the damping positive, even when c has a very large value. In particular, regardless of the choice of c, we can now give the damping such a value as to render the system aperiodic, this being the most favourable condition for our purpose. In order to facilitate this we have kept the delay time τ as small as possible, inter alia by arranging the aforementioned rectifying circuits in such a way that they can have a short RC-time without causing any excessive ripple in the direct voltage supply (cf. fig. 6). As may be seen from eq. (2), the delay time τ_1 of the speed control has no effect upon the damping.

A high value of τ_1, as also of d, however has an adverse effect upon the inertia term. Its effect is adverse because the aperiodic response of the system should be such that the final value is reached at a sufficiently slow rate. At the beginning of this article it has been stated that the edge of the picture may take the shape of a wave with a minimum wavelength of 4 cm. This means that a phase displacement of the maximum value may be made to disappear (or brought to the new final value) only after 2 cm picture length, that is to say after about $^1/_2$ second. The inertia term in eq. (2) must be made sufficiently large to bring this about. It was found, however, that this requirement could be easily met, independently of the reduction factors and delay times of the two controls, by increasing the moment of inertia I with the aid of a flywheel on the motor shaft.

Finally we give in *fig. 8* a simplified diagram of the whole synchronization system, showing how the phase control and the speed control have been combined. Each of the controls works on a separate output valve yielding a contribution to the field current of the receiving motor 5). A resistor (R_a) shunted across the output valves ensures that there is a certain field even when the valves are cut off, thus allowing for automatic starting (without the armature current reaching an excessive value); this also ensures that in the event of failure of the control voltages the motor will not race. The adjustment of the desired aperiodic damping of the phase variations is brought about not by the choice of the coefficient d, that is the amplification of the speed regulation, but by the choice of c, the amplification of the phase control, with the aid of the potentiometer P. The motive for this lies in the fact that the reserve in the sensitivity of the phase control in the synchronization circuit proved to be much greater than that in the speed control.

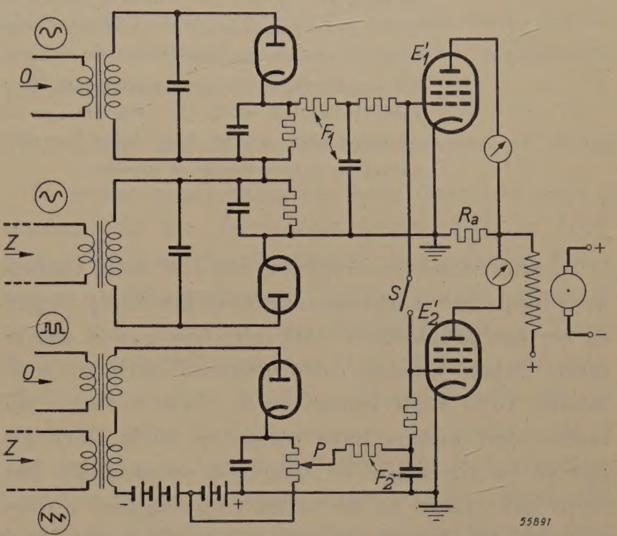

Fig. 8. Combination of phase control and speed regulation. The voltages O are derived from the receiver motor, Z from the transmitter motor. By means of the potentiometer T the "amplification" of the phase regulation — which provides sufficient reserve for the sensitivity required — can be purposely reduced so as to obtain the desired aperiodical damping effect of the speed regulation upon the phase variations. This adjustment applies only for a given ratio of the anode currents of the two output valves. This ratio is given the right value (in this case the value 1) by slightly varying the grid bias of E_1', which is supplied by the rectifier G_2' (see fig. 4). This is checked with the aid of the two milliammeters in the anode circuits. By closing the switch S the effects of the phase regulation voltage and the speed regulation voltage upon the two output valves are "mixed"; this, however, is not essential for the working of the synchronization system. F_1 and F_2 are smoothing filters.

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5) Actually the two controls are "mixed", for although two regulating valves are used we apply to the grid of each valve a voltage composed of fractions of the two regulating voltages.

THE ILLUMINATION OF COAL-MINES AND THE ATTENDANT RISK OF EXPLOSIONS

by G. D. RIECK.

622.476.4:622.81

When fire-damp occurs in coal-mines and it is mixed with air it is apt to cause explosions as soon as it comes into contact with a hot object. This article first deals with the steps that can be taken against this danger when oil lamps and electric incandescent lamps are used. It is then discussed how the danger of explosion may arise when fluorescent lamps are employed. An adequate safeguard is to apply a type of circuit in which the fluorescent lamp is ignited while the cathodes are cold. In mines the lamps are not switched on and off frequently, so that there is no fear of the life of the lamps being very much shortened owing to this method of ignition.

In coal-mines, especially at the coal face, there is very frequently a certain amount of fire-damp. Mixed in certain proportions with air, this gas is apt to ignite when coming into contact with a body heated to a high temperature. That is why the lamps that miners have to use in their work are apt to be the cause of disastrous explosions. The steps that have to be taken to safeguard miners against this danger are related to the method of illumination applied.

The risk of explosions was particularly great when the miners used an unprotected oil lamp, and it was therefore of great importance when Davy invented his safety-lamp in 1815. In this lamp the oil flame is screened off by a fine-meshed metal gauze. When a combustible mixture of fire-damp and air penetrates to the flame it ignites, but the hot combustion gases cannot spread the combustion through the gauze to the surroundings, filled with the same combustible gas. The gauze must have a certain mesh width (a maximum of 0.05 cm is prescribed). Provided the rate of flow of the combustion gases is not too high no explosion can then take place, although the gauze may be heated even to glowing temperature. Another advantage of this safety lamp is that the colour of the flame betrays the presence of mine gas.

Incandescent lamps for the illumination of mines

At the turn of the last century electricity began to be used more and more for the lighting of mines. At first only incandescent lamps were employed. An incandescent lamp can only cause an explosion when its bulb bursts in an atmosphere dangerously charged with fire-damp and the filament happens to keep glowing. This risk is not a great one and is most likely to occur close to where the coal is being hewn out, for example owing to a flying piece of coal hitting the lamp.

The breaking of a bulb, whereby the filament or at least a part of it is left glowing in a space charged with combustible fire-damp, need not always cause an explosion ¹). The conditions under which an explosion occurs or can be prevented have been investigated with an apparatus designed by Fripiat and represented diagrammatically in *fig. 1*. By this means it has been found, for instance, that the risk of explosion can be avoided by using a thin filament in the lamp ²). But lamps with a thin filament easily break and naturally do not yield much light. The risk of explosion can also be reduced by lowering the incandescent temperature of the wire, but then the efficiency of the lamp is much less. This is a great objection particularly for mine lamps of low power, which are fed from accumulator batteries. To get a reasonable efficiency a high incandescent temperature is required and the filament must be made of a metal having a very high melting point, such as tungsten. This material, however, oxidizes very quickly when exposed at a high temperature to air or mixtures of air and fire-damp, then flaming up and burning through; although the mixture of air and methane may not be immediately ignited by the hot wire (its incandescent temperature is at least 2000 °C) the flame arising when the wire burns through will most certainly set fire to the mixture.

When considering the risk of explosion one has to differentiate between the portable lamps (battery-fed), which the miners carry in their hand or have fixed to their caps, and the larger lamps used as fixed light points connected to the mains. The latter are usually located in well-ventilated galleries where there is so little danger that as a rule no special precautions need be taken. The portable

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¹) Extensive experiments have been carried out in this connection, inter alia, by J. Fripiat, See "Annalen der Mijnen België", 64, 105-144, 1943.

²) A platinum wire less than 0.1 mm in diameter caused to fuse by means of an electric current in a mixture of air and methane does not ignite the gas.

lamps are usually protected with an outer envelope (a wire-gauze screen would intercept too much light) and, moreover, in many cases mechanical safeguards are applied so that in the event of the outer bulb being broken a spring throws the lamp out of the fitting, thereby precluding any possibility of current continuing to flow through the filament. These measures, however, do not offer absolute safety, there still being a certain, though small risk, it being possible for a hole to be made right through the outer envelope and the inner bulb of the lamp without the safety mechanism being brought into operation, because this only reacts when the outer envelope is almost completely destroyed. The tungsten filament would then continue to glow at a high temperature and any mixture of air and fire-damp coming into contact with it would ignite; this is an exceptional risk of only sporadic occurrence. There is also even a possibility of the after-glow of the filament in the thrown-out lamp igniting the gas mixture, but in by far the majority of cases the safety mechanism will avert the risk of explosion.

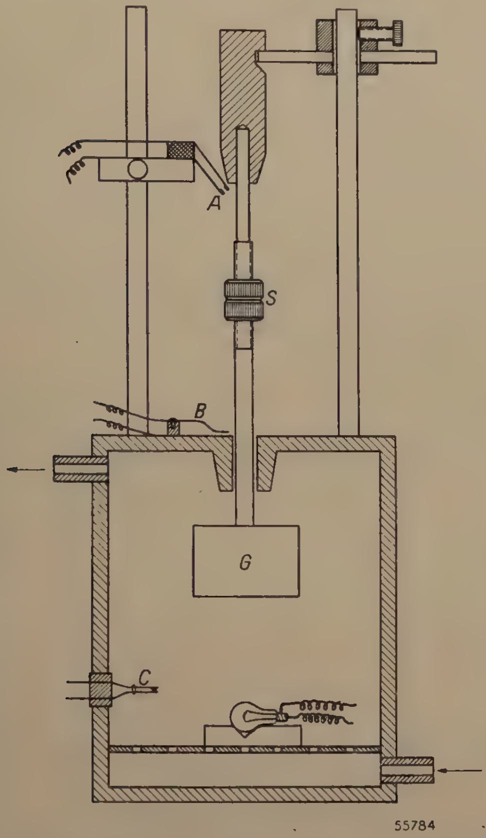

Fig. 1. Diagrammatic representation of an apparatus for testing mine lamps in respect to the danger of explosion. Such an apparatus was first designed by J. Fripiat. It consists of a metal explosion vessel with one wall of paper or cellophane. A mixture of air and methane flows into the vessel at the bottom and leaves it again through an opening at the top. The lamp to be tested is laid on a block resting on the bottom of the vessel. A metal weight *G* can be dropped onto the lamp to destroy the bulb. The nut *S* is first so adjusted that the filament of the lamp is not destroyed by the falling weight. Various electrical contacts are provided so as to be able to study the phenomenon of the combustion when that arises. The falling hammer closes a contact at *A* which starts the horizontal time-base of a cathode-ray oscillograph, and at *B* a contact is closed the moment that the lamp bulb is smashed, thereby bringing about a vertical deflection in the oscillogram. The moment of ignition of the gas mixture is recorded with a constant delay owing to the flame igniting a small piece of gun-cotton at *C*, thereby breaking the contact between two interconnected resilient plates. In many cases there is no need for these recording apparatuses when the object of the test is merely to ascertain whether ignition takes place or not.

Fluorescent lamps for the illumination of mines

In recent times there has been more and more a tendency to use fluorescent lamps (TL or MCF lamps) [3]) for the illumination of mine galleries. These can be used for permanent lighting points and small types can serve as portable light sources. These fluorescent lamps with their high efficiency and relatively low surface brightness have certain advantages for mine lighting because the black walls of the galleries reflect but little light and therefore call for rather powerful light sources. With the low level of brightness found in mine galleries a light source with a high brightness may easily cause troublesome glare. In this article we shall not go into the question whether the advantages of TL lamps outweigh the disadvantages, such as larger dimensions and higher initial cost. Here we are only concerned with the question in how far these lamps involve risk of explosions and what countermeasures have to be taken.

The fluorescent lamp requires stabilizing and an automatic starter. There are various types of starters in use: resistance-bimetal, glow-bimetal and electromagnetic starters [4]). For the problem that we are studying it does not matter much with what type of starter the fluorescent lamp is fitted. Let us assume that it is a glow-bimetal starter; the circuit normally used for this is represented in *fig. 2*. As soon as the mains are switched on a glow discharge takes place in the starter and heats the bimetal, with the result that after a time the contact is automatically closed. The choke is then short-circuited across the cathodes and the short-circuit current heats them. The glow discharge now being extinguished, the bimetal cools down, thereby opening the contact and cutting off the short-circuit current. The resultant voltage impulse ignites the lamp, the cathodes of which have mean-while reached a high temperature. Should the lamp fail to ignite, the process is automatically repeated, the cathodes being brought to glowing temperature each time. Once the lamp has been ignited no other current flows through the cathodes than that of the gas-discharge. This discharge ensures that the cathodes are kept at a temperature high enough for the lamp to burn.

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[3]) A. A. Kruithof, Tubular luminescence lamps for general illumination, Philips Techn. Rev. **6**, 65-73, 1941.

[4]) These types have been described in an article by Th. Hehenkamp, A rapid-action starter switch for fluorescent lamps, Philips Techn. Rev. **10**, 141-149, 1948 (No. 5).

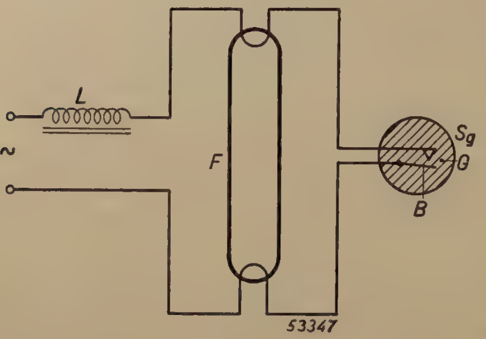

Fig. 2. The normal circuit of the glow-bimetal starter (*Sr*) of a fluorescent lamp (*F*). *G* = gas filling, in which a glow discharge may take place which heats the bimetal (*B*). *L* = ballast choke.

An explosion can only take place in the event of the glass tube of the fluorescent lamp breaking. In order to investigate how great the risk of explosion may be in such an event, we purposely broke TL lamps having the starting mechanism of fig. 2 in a Fripiat apparatus charged with a combustible mixture of air and fire-damp [5]). An explosion took place every time about 1 second after the breaking of the lamp. This is understandable when it is borne in mind that after the destruction of the lamp the starter still tries to ignite it. The short-circuit current will cause the rather thick wires of the cathodes to glow at a much higher temperature than that brought about by the normal discharge, and the glowing cathodes then cause the gas mixture to explode.

When the glass tube of a fluorescent lamp is broken one must indeed reckon with the possibility of the electrodes remaining intact. This is much more likely to be the case than with the so much smaller incandescent lamp, the filament of which, moreover, is much more fragile. To counteract the risk of explosion a construction is applied whereby a broken lamp is automatically thrown out of its fittings, so that no current can flow through the cathodes. Just as is the case with incandescent lamps, however, here again such a construction has the drawback that the tube of the lamp may be cracked or a piece of glass may be knocked out of it and leave a leak, without the lamp being ejected. Although the lamp is then extinguished, the very first time the starter tries to ignite it again the gas mixture in the lamp will explode.

The firing of the gas mixture in the tube need not necessarily lead to an explosion of the gas mixture in the surroundings. Tests have shown that a crack in the lamp or a very small hole (e.g. 0.5 mm in diameter) is not sufficient to allow of an outward propagation of the explosion taking place inside the tube; this may be compared to the effect of a metal gauze (as used in the Davy safety-lamp) in checking the spread of an explosion. In the event, however, of larger holes being made (e.g. 1 cm in diameter) and also if in consequence of the explosion inside the lamp the cracked tube falls to pieces, then the explosion is indeed spread outwards.

A better and more adequate method of avoiding explosions is to arrange for the lamp to be ignited while the cathodes are cold. This is possible when the circuit arrangement as sketched in *fig. 3* is applied, whereby the electrodes are not previously heated by a short-circuit current. In other cases this method of ignition is usually avoided in view of its attendant drawbacks. Obviously, if the electrodes are not preheated and no other measures are taken then as a rule a much higher voltage impulse is required to bring about the ignition, and in some cases this necessitates a special starter having a rapid current-breaking action. When small power lamps are used (25 or 40 W), however, the normal starter usually suffices.

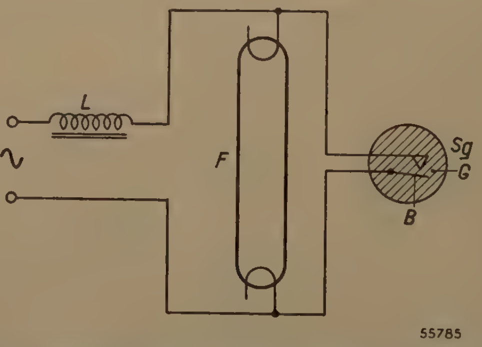

Fig. 3. A method of connecting the starter of a TL lamp which avoids the risk of explosion when the lamp is used in mines. With this arrangement the lamp is ignited with cold cathodes. For the meaning of the letters see fig. 2.

As a second objection it is mostly argued that with this method of ignition the electrodes suffer more and thus have a shorter life. For the lighting of mines, however, this is not of such great importance because the periods of operation are long and the lamp does not have to be switched on and off so frequently. Thus the advantages of this method of ignition most certainly outweigh the disadvantages.

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[5]) The tests were carried out with fluorescent lamps of special dimensions, so constructed as to fit in the testing apparatus.

We have carried out a number of tests with fluorescent lamps in a circuit according to the scheme of fig. 3. Experiments with a broken or leaking TL lamp fitted with electrodes like those of the TL 40 W lamp, whereby under these conditions no ignition takes place, did not give rise to any explosion, as was only to be expected considering that the cathodes remained cold. Such a lamp when ignited and placed in a Fripiat apparatus in a mixture of air and fire-damp and then broken to pieces did not cause any explosion either, the working temperature of the electrodes not being high enough to ignite the gas mixture after the discharge had been extinguished. Consequently a TL lamp used in the manner described above may be considered safe under any circumstances.

Care must be taken that the starter mechanism itself does not involve the risk of explosion. A resistance-bimetal starter, for instance, will certainly cause an explosion if the heating wire should be switched on while in combustible surroundings. In our tests an electromagnetic starter working normally while in contact with a mixture of air and fire-damp did not result in any explosion spreading to the surroundings. When, however, the ignition failed (for instance owing to a defect in the lamp) and the starter therefore continued to operate for an abnormal length of time, there was indeed a risk of explosion. No serious troubles, however, are to be expected from starters, it always being possible to place these in metal containers, and it is most unlikely that such a container would break without throwing the starter mechanism out of order.

Finally we would point out that our experiments were carried out with mixtures of air and methane of different compositions. We started with natural fire-damp of the composition: about 39% CH_4, 4% CO_2, 4% O_2, 9% H_2, 44% N_2 and 0.2% higher hydrocarbons. To produce a highly explosive mixture of this gas with air a quantity of O_2 was first added to the fire-damp such as to give the same ratio of oxygen to nitrogen as that in air, after which this gas mixture was diluted with air to a mixture containing 7-8% CH_4. It appeared that mixtures of air with this percentage of methane are the most dangerous as regards ignition temperature [6]), rate of propagation of the flame and suchlike. Tests were also carried out with a technical methane containing unsaturated hydrocarbons, instead of the natural fire-damp. These admixtures increase the combustibility of the mixtures of air and methane, but in no single instance did an explosion take place when the lamp was ignited with cold cathodes.

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[6]) The ignition temperature of these mixtures is about 650 °C.

A DIRECT-READING DYNAMIC ELECTROMETER

by J. van HENGEL and W. J. OOSTERKAMP. 621.317.723.082.742:621.386.82

A description is given of a direct reading d.c. electrometer. It is based on the dynamic principle, the direct voltage being converted into an alternating voltage with the aid of a parallel-plate capacitor, the capacitance of which changes periodically as one plate is kept stationary while the other is kept in vibration. The alternating voltage is applied to an amplifier with conventional valves; the output voltage is rectified and the conducted direct voltage measured with a moving-coil meter. In order to keep the amplification constant, a high degree of negative feedback is applied by connecting the output d.c. voltage wholly or partly in opposition to the voltage to be measured. This makes it necessary to use a special rectifying circuit which is sensitive to the polarity of the a.c. input voltage. Due in part to the direct-voltage feedback, the (apparent) input resistance reaches extremely high values (more than 10^{15} Ω). Two instruments are discussed: a millivoltmeter (full deflection at 100 mV) for laboratory measurements, and a dosimeter for X-rays (in combination with an ionization chamber and a measuring resistor). The dosimeter has a scale calibrated in r/min and covers a very wide measuring range, from $4 \cdot 10^{-7}$ r/sec (fraction of the tolerance dose) to 200 r/sec (contact therapy). For the calibration a standard-current generator has been designed which supplies a calibrated current of 10^{-9} A.

In many investigations in the field of physics and of electrical engineering the tension of direct-voltage sources which have a high internal resistance have to be measured. It is then essential that the voltmeter used should draw as little current as possible, so that the difference between the e.m.f. of the voltage source and the measured voltage under load is kept as small as possible.

Sometimes, when small direct currents are to be determined, the voltage produced across a high-ohmic resistor through which the currents flow is measured. Such cases occur not only in physical investigations but also, as will be shown farther on, in the daily routine of X-ray therapy. In these cases, too, one must have a voltmeter with very high resistance, since the meter bridges the measuring resistor and thus sets an upper limit to the resistance formed by the parallel connection of the two, thus limiting the sensitivity of the circuit.

In the cases mentioned one may often advantageously use an electrometer. This is an instrument in which the deflection is brought about by the electrostatic effect of charges and which therefore, in principle, consumes no current at all (apart from the charging current required to give the instrument a deflection, and the leakage current due to the finite insulation resistance).

A special form of electrometer is a triode — or, more generally, an electronic valve with a control grid — so arranged that no grid current flows through it. The word "no" is not to be taken too strictly, since owing to various causes the grid current is not generally absolutely zero; such causes may be imperfect insulation between the grid and other electrodes, ionic current flowing to the grid, or thermionic and photo emission from the grid. The methods applied in the special triode for use as an electrometer (type 4060) have already been described in this journal [1]). Within a wide range of conditions the grid current of this triode can be reduced to less than 10^{-14} A.

Here an electrometer will be described which, at full deflection, has a current consumption even 10^2 to 10^3 times smaller, so that the internal resistance of the source of the voltage to be measured, or the resistor through which the current to be measured flows, may be a corresponding factor larger. Moreover, this instrument has the advantage that, as opposed to some other electrometers, it gives a direct reading, whilst furthermore no special valves are required, only normal pentodes and rectifying valves being employed.

Principle of the dynamic electrometer

Our instrument is a further development of the dynamic electrometer described by Dorsman [2]), the principle of which is as follows. The direct voltage to be measured is applied via a high-ohmic resistor to a capacitor one of the plates of which is kept in vibration (for instance, with the aid of a loudspeaker system — see *fig. 1* — and a valve oscillator), so that the capacitance varies periodically. Just as in the case of a condenser-microphone, an alternating voltage is produced across the capacitor which at a given amplitude of vibration is proportional to the direct voltage applied. This alternating voltage is led via a coupling capacitor (C_2, *fig. 2*) to a normal a.c. pentode amplifier. The input resistance of the electrometer is mainly governed by the insulation of the coupling capacitor and that of the vibrating capacitor. With amber, polystyrene and similar substances as insulating material (and if desired as dielectric in the coupling capacitor) both capacitors can be given a much higher insulation resistance than is possible with an electrometer triode.

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[1]) H. van Suchtelen, The electrometer triode and its applications, Philips Techn. Rev. **5**, 54-59, 1940.

[2]) C. Dorsman, A p_H-meter with a very high input resistance, Philips Techn. Rev. **7**, 24-32, 1942; see also P. H. Clay, thesis, Amsterdam 1942; H. den Hartog and F. A. Muller, Physica **10**, 167-172, 1943 and **11**, 161-166, 1943; Hl. de Vries, Physica **13**, 449-452, 1947; H. Palensky, R. K. Swank and R. Grenchik, Rev. sci. Instr. **18**, 298-314, 1947.

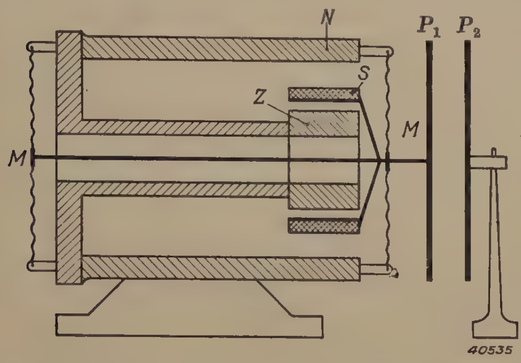

Fig. 1. Vibrating capacitor in cross section. N, Z = poles of a permanent magnet, S = coil of the electrodynamic driving system, M = diaphragm springs, P_1 = vibrating plate, P_2 = stationary plate of the capacitor.

The electrometer described by Dorsman was designed for measuring the potential of a "glass electrode", from which potential the acidity (p_H) of a solution can be derived. The method of measuring is a zero method: the voltage to be measured is applied to the input of the electrometer in opposition to a reference voltage derived from a compensator, the latter voltage being so adjusted that the alternating voltage at the output of the amplifier is zero; this is observed on a cathode-ray indicator connected to the output. When the output voltage is zero the two d.c. input voltages are equal, so that the unknown potential may be read from the compensator calibrations.

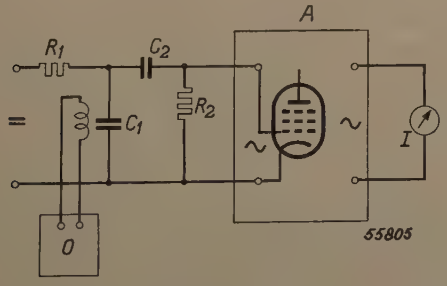

Fig. 2. Circuit diagram of the dynamic electrometer. C_1 = vibrating capacitor, the driving coil of which is connected to an oscillator O; R_1 = high-ohmic series resistor counteracting changes in the charge of C_1 when the system is vibrating, C_2 = coupling capacitor, R_2 = grid leak resistor, A = alternating voltage amplifier, I = indicator.

The zero method is suitable for the accurate measurement of very low voltages and has, moreover, the advantage that variations in the amplification or in the amplitude of the vibrating capacitor (due, for instance, to mains voltage fluctuations) only affect the precision of the adjustment and not the reading.

For many purposes, however, a direct-reading instrument is to be preferred, where no adjustments have to be made for every measurement, so that one's hands are left free and, moreover, the work can be done more quickly. It is then obvious that there should be connected to the output of the electrometer amplifier a measuring instrument provided with a scale calibrated in the units of the quantity to be measured, for instance mV.

Of course the favourable features of the zero method were to be retained as far as possible, and therefore the amplifier had to be of such a type that its amplification factor is reasonably constant, both in the event of mains voltage fluctuations and in the case of a variation of the characteristics of the valves. This requirement of constancy in the amplification has made it necessary to introduce drastic modifications of the electrometer amplifier used for p_H-measuring before it was possible to realize a direct-reading instrument.

Following these lines we have so far made two types of direct-reading dynamic electrometers for laboratory work, one as a mV-meter with very high impedance for general laboratory use, the other provided with a scale in röntgens per minute for measuring X-ray doses (with the aid of an ionization chamber). The circuit arrangements of both these types are mainly the same.

Direct-voltage feedback

The desired constancy of the amplifying factor is obtained by applying feedback [3]) to a high degree, by returning the output voltage E_2 (*fig. 3*) either wholly or in part to the input and there connecting it in opposition to the voltage to be measured, E_1. The output voltage must therefore be a direct voltage (which can thus be measured with a moving-coil meter); it is obtained with the aid of a rectifier connected to the output of the amplifier. Thus the feedback is a direct-voltage feedback.

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[3]) B. D. H. Tellegen, Inverse feed-back, Philips Techn. Rev. 2, 289-294, 1937.

Across the vibrating capacitor is the direct voltage $E_1—E_2$. The vibration sets up a proportional alternating voltage v_1.:

$$v_1 = a_1\,(E_1—E_2),$$

which is amplified by a factor a_2 to the a.c. output voltage v_2:

$$v_2 = a_2\,v_1 = a_1a_2\,(E_1—E_2).$$

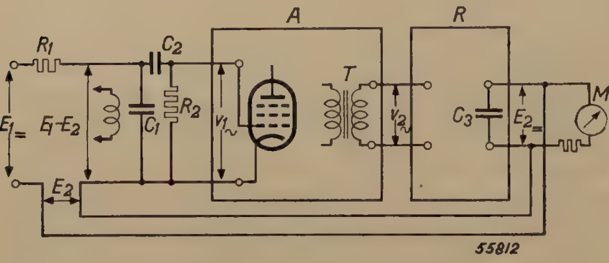

Fig. 3. Circuit diagram of the direct-voltage negative feedback system applied in our electrometer. E_1 = direct voltage to be measured; E_2 = d.c. output voltage; v_1, v_2 = a.c. input and output voltage, respectively of the amplifier A; T = transformer; R = rectifier with smoothing capacitor C_3; and M = moving-coil voltmeter. The other letters have the same meaning as in fig. 2.

From this the rectifier produces the direct voltage E_2 proportional to the amplitude of v_2 (proportionality factor a_3):

$$E_2 = a_1a_2a_3\,(E_1—E_2).$$

Thus we have for the ratio p between the d.c. output voltage E_2 and the direct voltage to be measured E_1:

$$\frac{E_2}{E_1} = p = \frac{a_1a_2a_3}{1 + a_1a_2a_3} = \frac{A}{1 + A}, \quad . \; . \quad (1)$$

when $A = a_1a_2a_3$.

The factors a_1, a_2 and a_3 depend more or less upon the properties of the oscillating, amplifying and rectifying valves respectively and upon the supply voltages (and thus the mains voltage). If however, it is arranged that $a_1a_2a_3$ is always large with respect to unity, then $p \approx 1$ and therefore practically constant, in spite of fluctuations in the values of a_1, a_2 and a_3. Now a_1 and a_3 are both smaller than unity, but a_2 can be made so large that in fact $A = a_1a_2a_3 \gg 1$. This requirement has been met in both forms of construction of our electrometer.

A limit is set to the raising of the value of A owing to the fact that at a higher amplification the system would become unstable, that is to say it would oscillate at a frequency depending upon the time constants of the circuits contained in the system. The larger the time constants, the more the amplification can be raised, but the adjustment time of the instrument becomes longer.

With the method of feedback applied the direct voltage across the vibrating capacitor is not E_1 but only $E_1—E_2$, which may well be 30 times smaller than E_1. This means that the leakage current flowing across the capacitors C_1 and C_2 at a given insulation resistance is reduced. It is partly due to this fact that the (apparent) resistance between the input terminals is exceptionally high.

As regards the a.c. losses in the resistors R_1 and R_2 (and the resistor R_5 to be discussed farther on), it must be noted that these are supplied by the oscillator driving the vibrating capacitor. These losses, therefore, do not form any load upon the voltage source to be measured.

Rectification

If the rectifier (R, fig. 3) were arranged according to one of the conventional methods, where the direct voltage obtained depends only upon the amplitude and not upon the phase of the a.c. input voltage, an unstable situation would arise, as may be seen from the following numerical example.

Let A = 30, thus, according to eq. (1), p = 30/31, and let E_1 be 100 mV, thus E_2 = (30/31)·100 = 97 mV. (The scale of the moving-coil meter to which E_2 is applied is such that the instrument indicates $E_2/p = E_1$, in this case 100 mV). $E_1—E_2$ = 3 mV.

Let us suppose, further, that E_1 suddenly drops to 94 mV. Since E_2 cannot change quickly, $E_1—E_2$ becomes at first —3 mV, in consequence of which v_1 and v_2 maintain the same amplitude but change polarity, which amounts to a phase shift of 180°. An ordinary rectifying system, however, is insensitive to this change and E_2 is kept at 97 mV. The situation is now unstable: any further drop of E_1 is accompanied by a rise in the absolute value of $E_1—E_2$, and also in the output voltage E_2. The negative feedback is turned into a positive feedback.

Besides this hypothetical experiment, any normal switching-on leads to instability, since E_2 builds up in an oscillatory way, so that there will be moments when $E_1—E_2$ is negative and the pointer of the instrument swings beyond the end of the scale.

To avoid this instability a rectifying system has been designed in which, when v_2 changes 180° in phase, E_2 tends to drop and to restore the proper polarity of $E_1—E_2$. When E_2 has reached p times the new value E_1 — in the example just given, therefore, (30/31)·94 = 91 mV — equilibrium is again obtained, $E_1—E_2$ again having practically its original value (+ 3 mV). At this lower value of E_2 (91 mV) the meter indicates the exact value of E_1 (94 mV).

Fig. 4 represents the rectifying system in question. The polarity of v_2 is compared with that of the alternating voltage v_3 derived from the oscillator driving the vibrating capacitor. Special measures (to which we shall refer later) have been taken to ensure that the phase difference between v_2 and v_3 can only be either 0° or 180°.

This systems functions in the following way.

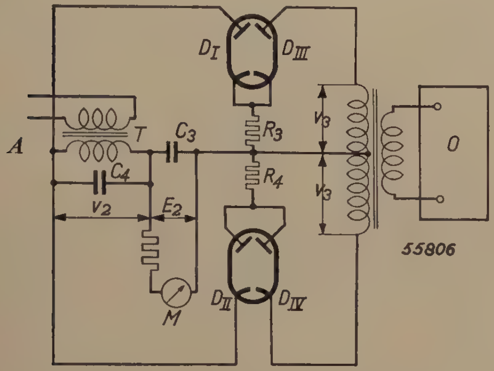

Fig. 4. Circuit of the phase-sensitive rectifier. T = output transformer of the amplifier; C_4 capacitor with which T is tuned; C_3 = smoothing capacitor; D_I-D_{III} and D_{II}-D_{IV} = double diodes (EB 4); O = oscillator also driving the vibrating capacitor.

For alternate half cycles [4] the alternating voltage v_0 causes current to flow through the diodes D_{III} and D_{IV} and the resistors R_3 and R_4. During these half cycles voltages are developed across these resistors which, apart from voltage losses, are equal to v_3. The resistors also form part of the circuits via which the smoothing capacitor C_3, across which the output voltage E_2 is developed, either receives charge from the transformer T (via the diode D_I) or is discharged (via the diode D_{II}). When the phase shift between v_2 and v_3 becomes 180° then — as will be explained below — a change takes place in the ratio of the quantities of the charge fed to and drawn from C_3 per cycle, thus changing E_2 in the desired direction.

Let us first consider the situation where the direct voltage across the vibrating capacitor has the normal polarity (E_1—$E_2 > 0$). The system must be then so adjusted that v_2 has the polarity at which A is positive with respect to the terminal of T connected to C_3 (fig. 4) during the half cycles in which D_{III} and D_{IV} are non-conducting; within these half cycles the capacitor C_3, receives a charge via D_I (so long as $v_2 > E_2$) and gives off a charge via D_{II} (so long as $v_2 < E_2$) these valves not being blocked by the voltages v_3. During the other half cycles the diodes D_{III} and D_{IV} are conducting and, as we have seen, across each of the resistors R_3 and R_4 there lies the voltage v_3. This voltage is chosen high enough to block the diodes D_I and D_{II}; thus the voltage v_2 does not cause any current to flow. As a result the voltage E_2 adjusts itself in such a way that the capacitor C_3 receives (via D_I) per cycle just as much charge as it gives off (via D_{II} and via the meter connected to C_3).

We shall now consider the case where E_1—$E_2 < 0$. The polarity of v_2 at which A is positive with respect to the terminal of T connected to C_3 then occurs during the half cycles in which D_{III} and D_{IV} are conducting, and consequently — just as was the case before — D_I and D_{II} are blocked. During the other half cycles (D_{III} and D_{IV} non-conducting) no charge can flow to C_3 via D_I (v_2 has the wrong polarity for that) but charge can flow from C_3 through D_{II}. Thus the case in question (E_1—$E_2 < 0$) leads to a drop in E_2, as was desired.

With this system two valves suffice if double diodes (type EB4) are used as indicated in fig. 4.

Any voltage superimposed upon v_2 with a frequency different from that of v_2 and v_3 as a rule contributes little or nothing towards E_2. In other words, the rectification is selective, so that an interfering voltage with the mains frequency or resulting from "microphonic effect" will have little influence. The selectivity is still further increased by tuning the transformer T (fig. 4) with a parallel capacitor (C_4) to the frequency of vibration (125 c/s, midway between two harmonics of 50 c/s, so chosen as to avoid trouble from these harmonics). Although this circuit has so much damping that the voltage gain is of little significance, harmonics of v_2 are effectively suppressed.

Amplifier

From the explanation of the functioning of the rectifying system it will be clear that changes in the phase angle between v_2 and v_3 will affect the amplitude of E_2. When, in the amplifier, circuits are used with little damping and tuned to the frequency of the vibration — as is the case in the amplifier designed for p_H-measurements (see footnote [2])) — then the phase of v_2 changes considerably with small variations in the frequency of vibration or in the event of a slight detuning. In order to avoid this we have not used any tuned circuits in the amplifier for our electrometer (the circuit T-C_4 previously mentioned — fig. 4 — is so strongly damped as not to endanger phase stability). We employ a resistance-coupled amplifier (*fig. 5*) with two stages. The value of the coupling capacitors C_5 and C_6 has been so chosen that, in combination with the resistors R_5 and R_6 respectively, they bring about an appropriate phase shift such as to bring

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[4]) For the sake of convenience we speak of "half cycles" although actually the intervals referred to differ slightly from half a cycle. We shall not go into this because it is of no consequence for our arguments.

v_2 exactly in phase with the reference voltage v_0 (fig. 4), as is desired for the rectification.

The dimensions of the circuit C_2-R_2-C_5-R_5 are such that voltages having the frequency of vibration are little attenuated in contrast with interference voltages derived from the first or second harmonics of the mains frequency. Thus the overall selectivity is improved. As an illustration of the strong filter action obtained in this manner it can be stated that an interfering voltage of 200 mV r.m.s., 50 c/s, superimposed upon an input direct voltage E_1 = 10 mV, does not affect the meter reading.

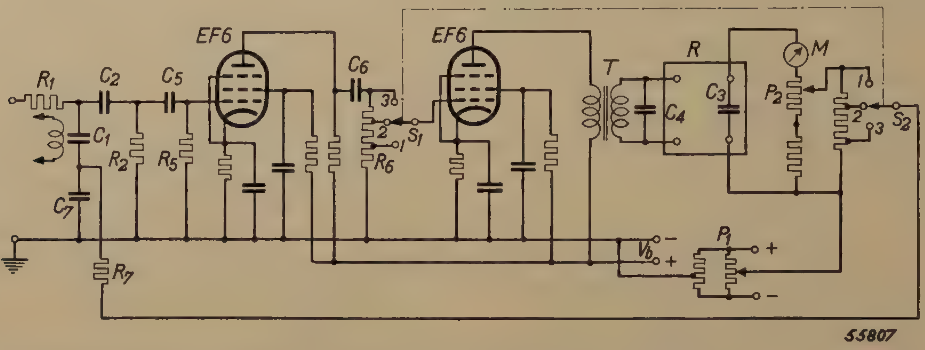

Fig. 5. More detailed diagram of the direct-reading electrometer (cf. fig. 3). R_5, C_5 and R_6, C_6 = circuits for bringing v_2 into phase with v_3; R_7, C_7 = smoothing circuit via which the d.c. output voltage (or a part of it) is returned to the input; S_1, S_2 = switches for varying the sensitivity: in position *1* the full deflection is at 1000 mV input voltage, in position *2* at 300 mV, in position *3* at 100 mV; P_1 = potentiometer for zero point correction; P_2 = potentiometer for scale correction; V_b = supply voltage. The other letters have the meaning given in figs 2 and 3.

Fig. 6 shows how little the reading is affected by mains voltage fluctations, due to the method of negative feedback applied; a variation in the mains voltage from the nominal value (100%) to 90% or 110% gives at most a difference of 1% on the meter.

Further details of the electrometer

The electrometer was designed for 100 mV at full deflection, corresponding to an accuracy within about 1 mV. This relatively low voltage-sensitivity enhances the reliability of the instrument and has, moreover, the advantage that as a rule no trouble is experienced from statistical current fluctuations. With this degree of sensitivity the direct voltage across the vibrating capacitor at the full deflection amounts to about 3 mV.

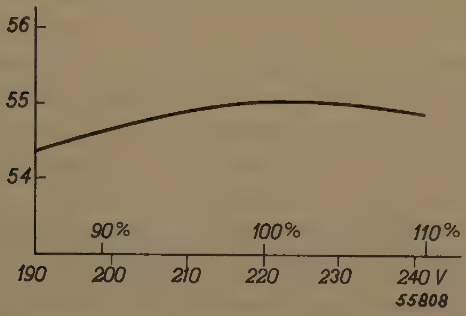

Fig. 6. Variation of the deflection (nominally, e.g., 55 scale divisions) as a function of the mains voltage fluctuations. If the mains voltage (rated value 220 V) fluctuates between 90% and 110% the variation in deflection is less than 0.5 division.

In cases where voltages higher than 100 mV have to be measured a switching device is used; this will be dealt with presently.

The reading is taken from a moving-coil meter giving full deflection when a current of 50 μA is flowing through the coil.

Zero adjustment is made in the first place, as far as the meter itself is concerned, with the normal zero adjustment of the pointer. In addition there is an electrical correction: by means of the potentiometer P_1 (fig. 5), connected to a small auxiliary direct voltage, any changes in the contact potential on the plates of the vibrating capacitor can be compensated.

The calibration of the meter can be corrected with the aid of the potentiometer P_2.

In the types already constructed the resistance between the input terminals of the electrometer is more than $10^{15}\,\Omega$ and the input capacitance about 40 pF.

Still more favourable values can be reached by applying the variation of the input circuit illustrated in *fig. 7* (so far this has not been normally applied). The insulation of the non-earthed input terminal is screened with a guard ring brought to the potential E_2. As we have already seen in the case of the capacitors C_1 and C_2, owing to this measure the insulation is only loaded with the voltage E_1—E_2 = about $E_1/30$. As a consequence the apparant input resistance is greatly increased and the apparent input capacitance many times reduced.

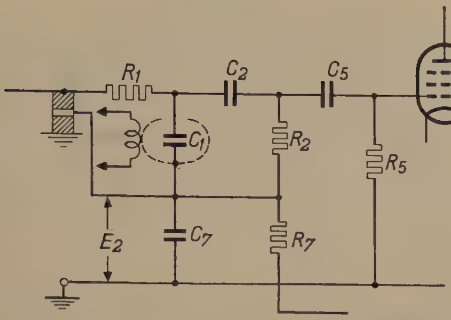

Fig. 7. Modification of the input part of the system according to fig. 5. In the insulator of the input terminal is a guard ring brought to the potential E_2, as is also the screening of C_1, so that the insulation at that spot is only loaded with the voltage E_1—E_2. This greatly increases the apparent input impedance and reduces the apparent input capacitance.

Practical application as laboratory voltmeter

Fig. 8 gives an illustration of the electrometer developed upon the principle described and intended for use as a laboratory instrument for taking measurements in cases where a very high input resistance is required. The circuit is that represented in fig. 5 (but without the switches S_1 and S_2) and in fig. 4. After what has already been said, this instrument needs no further explanation.

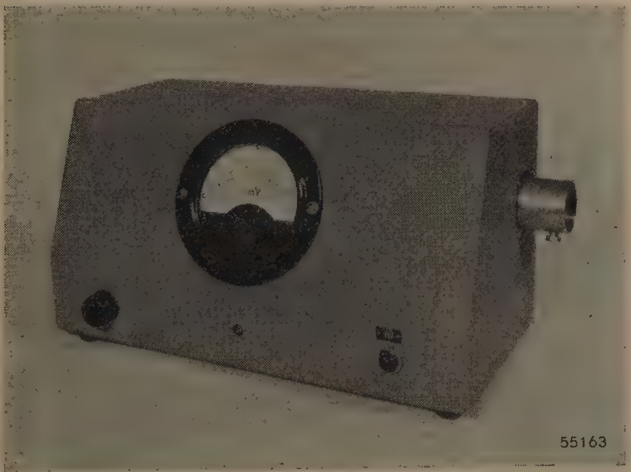

Fig. 8. Direct-reading dynamic electrometer for laboratory use. Full deflection at 100 mV input voltage. Input impedance $> 10^{15}\ \Omega$, input capacitance about 40 pF.

Practical application as dosimeter for X-ray therapy

The dosage of X-ray or gamma irradiation is expressed in terms of the röntgen unit (r), which is based upon the ionizing action of the rays and is defined as follows (Chicago 1937):

"The roentgen shall be that quantity of X or gamma radiation such that the associated corpuscular emission per 0.001293 gram of air produces, in air, ions carrying 1 e.s.u. of quantity of electricity of either sign".

A well-known method of measuring X-ray doses which is of course based on this definition is the following: an ionization chamber — a vessel containing a given quantity of air (or some other gas) and two electrodes between which a direct voltage is applied — is exposed to the rays, which ionize the gas and thus make it partially conducting. A current therefore flows between the electrodes. The potential difference should be large enough to cause the current of ions to reach its saturation value but must remain below the striking voltage. A current intensity of 3.33×10^{-10} A per cm^3 air (of one atm. and 0° C) in the ionisation chamber corresponds to 1 r/sec.

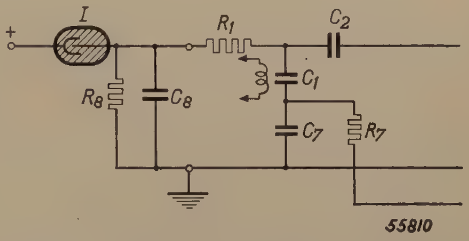

Fig. 9. Input circuit of the electrometer connected to an ionization chamber I with measuring resistor R_8 and smoothing capacitor C_8, for measuring the dose rate of X-ray irradiation. The other letters have the meaning given in fig. 5.

In order to measure the current it is passed through a high-ohmic resistor (R_8, *fig. 9*) and the voltage developed across it is measured. In view of the very weak current an electrometer is required for this purpose, and, given its exceptionally low current consumption, the electrometer described is particularly suitable.

Since the X-ray tube is fed with a more or less pulsating direct voltage (sometimes even with an alternating voltage) there is a certain "ripple" in the radiation, thus also in the ionization current thereby generated. A capacitor (C_8) shunted across the measuring resistor keeps the ripple voltage across the resistor sufficiently small.

Measuring range

The measuring range for which the dosimeter had to be designed is very wide: on the one hand it was desired to be able to measure the very large dose rates (doses per unit of time) which occur in contact therapy [5]) (up to 200 r/sec), while on the other hand it had to be possible to measure also fractions of the very much smaller dose rate corresponding to the amount of radiation that the tissues of the human body can bear without being damaged (e.g. 0.1 r/day, i.e. about 10^{-5} r/sec)[6]. These limits thus differ by a factor 10^9.

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[5]) See, e.g. H. A. G. Hazeu, J. M. Ledeboer and J. H. van der Tuuk, An X-ray apparatus for contact therapy, Philips Techn. Rev. 8, 8-15, 1946.

We shall now consider in what way the desired range in sensitivity of the instrument can be realized.

1) **Volume of the ionization chamber.** The current supplied by the ionization chamber is approximately proportional to its volume. A series of ionization chambers differing greatly in size therefore already permit an extensive measuring range to be covered. Of course the dimensions cannot be chosen arbitrarily, since one has to take into account the spatial distribution of the field of radiation in which the measurement is to be taken. For contact therapy we use a chamber of 25 mm^3, for deep therapy one of 6 cm^3 and for measuring tolerance doses one of 35 cm^3 or of 150 cm^3.

2) **Gas filling.** Usually the ionization chamber is in communication with the outside air and is thus filled with air under atmospheric pressure. Consequently for very accurate measurements corrections have to be applied for temperature and barometric pressure.

If it is decided to use a sealed ionization chamber then this can be filled with some other gas under a different pressure. If high sensitivity is required then a gas is chosen with a high atomic number and not too low pressure. The higher the atomic number of the gas, the greater is the absorption of the X-rays (at least if the rays are not very "hard") and thus also the greater the ionization brought about in the gas. In krypton (atomic number 36), for instance, the ionization current is about 400 times as large as that in air (atomic number of nitrogen 7 and of oxygen 8). An ionization chamber filled with krypton therefore renders good service for measuring very small dose rates. There is the disadvantage, however, that then the meter reading is dependent upon the hardness of the rays, so that a different calibrating constant has to be used for different qualities of radiation.

3) **Measuring resistance.** The higher the measuring resistance, the greater is the sensitivity. We use measuring resistors having a value between 50 MΩ and 2000 MΩ. Since the resistance between the input terminals of our electrometer is much higher, the measuring resistance can be further increased if required, but it is not easy to make resistors of such a high value which are sufficiently stable.

It is possible to do without the measuring resistor. Then the ionization current charges the input capacitance of the electrometer. The voltage reading follows the rising input voltage. The ionization current can be calculated from the values of the input capacitance and the time, measured by a stop watch, within which the needle covers a given distance on the scale. With this method it is advantageous to employ the system shown in fig. 7.

4) **Switching-over of the electrometer itself.** The measuring range can of course also be extended by giving the electrometer itself various degrees of sensitivity. In our instrument this has been done in the manner indicated in fig. 5: full deflection is obtained at an input voltage of 1000, 300 or 100 mV, according to whether the switches S_1 and S_2 are in the position *1*, *2* or *3*.

In the designing of this switching method we started from the consideration that the sensitivity of the moving-coil meter had to be kept unchanged, so that at the full deflection E_2 has the same value (about 970 mV) in all the three positions.

If, in order to get lower sensitivity, the moving-coil meter were shunted, then for the full deflection a higher alternating voltage would be required at the input of the rectifier and also a larger amplitude of the auxiliary alternating voltage v_3 supplied by the oscillator to the rectifier (fig. 5). The latter might result in a heavy loading of the oscillator; moreover the load would change when switching over to a different sensitivity, which might impair the constancy of the oscillator frequency and amplitude.

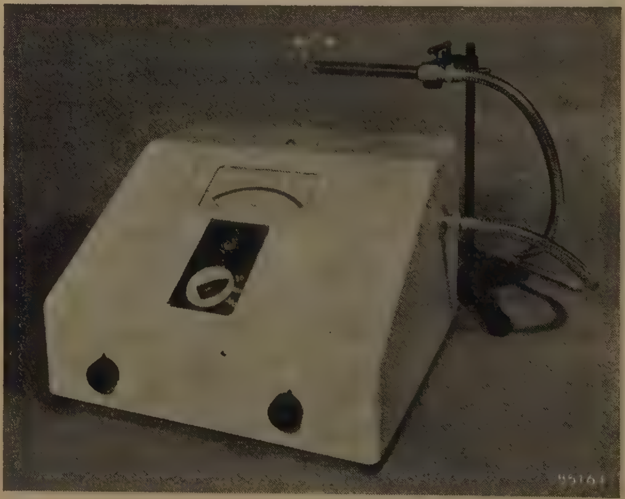

Fig. 10. Electrometer constructed as dosimeter, with scale in r/min. I = ionization chamber.

In the positions *2* and *3* (fig. 5) for max. 300 and 100 mV respectively only the part aE_2 of E_2 tapped off by S_2 is used as feedback voltage, the fraction being so chosen that it is always approximately equal to the same fraction p (≈ 0.97) of E_1. As may be seen from *table I*, in the positions *1* and *2* the direct voltage E_1—aE_2 across the vibrating capacitor is then respectively 10 and 3 times as high as in the position *3*; thus the amplification can be reduced to $^1/_{10}$ or $^1/_3$. This is brought about with the switch S_1. The amplification cannot reach such values as to give rise to danger of oscillation. Table I shows the voltages produced at full deflection.

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[6] It is customary to express these tolerance doses per unit of time in r/day or in r/sec, whereas the dose rates for therapeutical purposes are expressed in r/min. For the sake of clarity we use here only r/sec.

Table I. Direct voltages at full deflection in the three positions of the switches S_1 and S_2 (fig. 5). E_1 = voltage to be measured, E_2 = output voltage, αE_2 = negative feedback voltage, $E_1 - \alpha E_2$ = direct voltage across the vibrating capacitor.

Position		1	2	3	
E_1	=	1000	300	100	mV
E_2	≈	970	970	970	mV
αE_2	≈	970	291	97	mV
$E_1 - \alpha E_2$	≈	30	9	3	mV

For various fields of applications of X-rays a summary is given in *table II* of the possibilities mentioned sub (1), (2) and (3) which form a favourable combination, whereby $E_{1\max}$ may be made 1000 or 100 mV.

Table II. Table of the dose rates occuring in various fields of applied röntgenology, the ionization chambers used by us when measuring the doses, and the currents and voltage thereby obtaining.

Application	Dose rate to be measured	Ionization chamber		Ionization current	Measuring resistor R_8	E_1	E_1/sec
		volume	gas				
	r/sec	cm³		A	MΩ	mV	mV/sec
Contact therapy	0.5 - 250	0.025	air	$4\cdot10^{-12}$ - $2\cdot10^{-9}$	500-2000	8 - 1000	—
Deep therapy	0.1 - 5	6	air	$2\cdot10^{-10}$ - 10^{-8}	100	20 - 1000	—
Tolerance doses	$4\cdot10^{-7}$ - 10^{-5}	150	Kr *)	$2\cdot10^{-12}$ - $5\cdot10^{-11}$	2000	4 - 100	—
	10^{-6} - 10^{-5}	25	air	10^{-14} - 10^{-13}	∞ **)	—	0.6-6 **)

*) The data given on this line apply for about 80 kV across the X-ray tube.

**) Measured without measuring resistor and with the system according to fig. 7. The ionization current can be calculated from the rate at which E_1 increases and the value of the input capacitance.

With a given ionization chamber and a given measuring resistance the moving-coil meter can be calibrated directly in r/sec or r/min (*fig. 10*). The electrometer itself, as already stated, is insensitive to mains voltage fluctuations and variations in the mutual conductance of the amplifying valves, but changes in the measuring resistor (R_8, fig. 9) may cause errors. In order be able to check the instrument, including this resistor, at any time, we have constructed a reference current generator, an apparatus supplying a very small, constant, known current which, when passed through the measuring resistor, gives a characteristic reading of the electrometer.

Reference current generator

The system applied for this accessory is diagrammatically represented in *fig. 11*. Use has been made of the familiar property of a pentode that the anode current is practically independent, within wide limits, of the anode voltage. The anode current, however, is far too high for our purpose and therefore a capacitative current divider is used consisting of the capacitors C_9 (5 μF) and C_{10} (50 pF) bearing the relation of 10^5 : 1. The anode current is divided in the same ratio. The smaller of the partial currents passes through the measuring resistor R_8.

The current i passing through R_8 and C_{10} a period of time t after switching on is calculated as follows

$$i = \frac{C_{10}}{C_9 + C_{10}}\, I \left[1 - \exp\left(-\frac{C_9 + C_{10}}{C_9 C_{10} R_8}\, t\right)\right],$$

where I represents the anode current. When $C_9 \gg C_{10}$ one may write:

$$i \approx \frac{C_{10}}{C_9}\, I \left(1 - \exp\frac{-t}{C_{10} R_8}\right).$$

Even with R_8 = 2000 MΩ the time constant $C_{10}\, R_8$ is still only 0.1 sec, so that we very soon get

$$i \approx \frac{C_{10}}{C_9}\, I.$$

The process of calibration is as follows. When the switch S_3 is closed the anode current is adjusted to

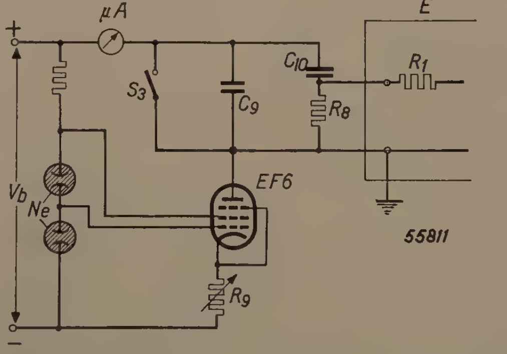

Fig. 11. Circuit diagram of a reference current generator supplying a current of 10^{-9} A for calibrating the dosimeter. V_b = supply voltage, Ne = neon tubes for stabilizing the voltage, R_9 = variable cathode resistor for adjusting the anode current of the pentode to 100 μA; C_9-C_{10} = capacitative current divider 10^5 : 1; R_8 = measuring resistor; E = electrometer.

100 μA with the aid of a variable cathode resistor. When S_3 is opened a current of $10^{-5} \times 100\ \mu$A $= 10^{-9}$ A flows through R_8, which should give a characteristic reading on the electrometer. Any deviation from the correct deflection can be corrected with the potentiometer P_2 (fig. 5). After about 15 sec the capacitors C_9 and C_{10} are so far charged as to cause the anode voltage to drop below the limit at which the anode current is constant; the current then rapidly diminishes. The interval of 15 seconds is long enough for the correction to be carried out.

SECONDARY EMISSION IN OUTPUT VALVES

by J. L. H. JONKER.

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Secondary electron emission is often an undesired phenomenon, which may, for instance, have a very adverse effect upon the functioning of tetrode amplifying valves. Known counter-measures are: 1) concentration of the space charge between screen grid and anode, 2) a third grid at low potential, 3) covering the anode with a layer of a substance from which secondary electrons cannot easily emerge. It is discussed why the means under 1 and 2, separately applied, frequently do not yield to the desired result, especially in the case of output valves As output valves, therefore, pentodes are usually employed in which advantage is also taken of the space charge. Output valves fed from an anode battery, however, operate with too small an anode current to be able to profit from the action of the space charge. Moreover, the supply voltage is low, and as a result the secondary emission consists of proportionately fewer slow, "real" secondary electrons and more rapid, reflected electrons. The latter are capable of passing the suppressor grid more easily. It appears that in such output pentodes the application of the third counter-measure leads to very much better characteristics and a higher output, with a given distortion. The new output pentode DL 41 for battery supply, with prepared anode, gives for instance with 10% distortion an output of approx. 260 mW, as against 200 mW for a similar valve with bare anode.

When electrons impinge upon the surface of a conductor or an insulator at a certain velocity some of them are reflected while the others penetrate into the material and transmit their energy to the electrons already in that material. In consequence some of the latter electrons, given a favourable direction of movement, emerge from the surface bombarded. This is the well-known phenomenon of **secondary emission** —by which one usually has in mind all the electrons coming from the bombarded surface, the emitted as well as the reflected ones. Several articles have already been written on this subject in this journal [1]).

Whereas in some cases good use can be made of this secondary emission, in others it is a most undesirable phenomenon and means have to be sought to suppress it as far as possible.

A familiar form of an electronic valve in which secondary emission may be highly injurious is the **tetrode**. The (primary) electrons which pass through the openings in the screen grid may strike the anode with considerable force and thereby liberate secondary electrons. When the tetrode is used as an amplifier, due to the anode load, the anode potential consists of an alternating voltage superimposed on the direct voltage. As a result the anode voltage may be temporarily lower than the constant screen-grid voltage. When this is the case the electrons are attracted towards the screen-grid, thereby considerably reinforcing the screen-grid current at the cost of the anode current i_a. In the characteristics representing i_a as a function of the anode voltage v_a irregularities then occur in the form saggings (*fig. 1*) which are apt to give rise to distortion. The anode current may not only drop to zero but may even be reversed in sign.

Means of suppressing secondary emission have already been discussed in this journal [2]) and consist mainly of the following measures:

1) concentration of the space charge between screen grid and anode,
2) the application of a third grid (at low potential),
3) covering the anode with a layer of a substance tending to prevent the emergence of electrons from the anode [3]).

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[1]) See for instance H. Bruining, Secondary electron emission, Philips Techn. Rev. 3, 80-86, 1938.

[2]) J. L. H. Jonker, Phenomena in amplifier valves caused by secondary emission, Philips Techn. Rev. 3, 211-216, 1938.

[3]) A similar effect is obtained when fins or vanes are provided perpendicular to the anode surface as mentioned in the article quoted in footnote [2]). Such fins are not considered here.

Hitherto it has not been usual to apply more than one of these counter-measures simultaneously in one particular type of valve, though there are cases where a combination of (1) and (2) has been applied. In this article we shall show how in a certain case a particularly favourable effect can be produced by the combination of (2) and (3), that is, the application of a third grid and at the same time a prepared anode.

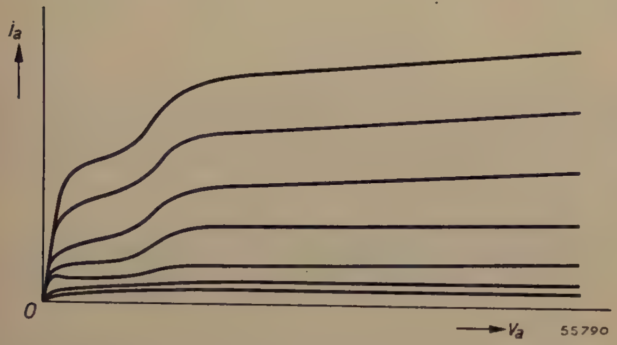

Fig. 1. Characteristics of a tetrode: anode current i_a as function of the anode voltage v_a at a constant value of the screen-grid voltage. Owing to secondary electron emission there is some sagging in the characteristics which gives rise to distortion.

First of all we shall consider briefly the measures referred to under (1) and (2), when we shall see the circumstances under which they fall short of their purpose.

Concentration of space charge

If a potential minimum is produced between screen grid and anode then, provided there is sufficient depth of the minimum, secondary emission is counteracted. A potential minimum can be obtained, for instance, by causing a concentration of the space charge between the said electrodes, this method often being used in tetrodes and, as we shall see, sometimes also in pentodes. Such a concentration may be brought about, for example, by giving the electrodes a certain shape and position, whereby the electrons are concentrated into one or more beams.

When one is dealing with output valves where it is of importance that the anode current and anode voltage should reach the largest possible amplitudes with the least possible distortion, the potential minimum should be present at widely divergent momentary values of that anode current and voltage. The meaning of this is illustrated in *fig. 2*, where the static characteristics of an output pentode ($i_a = \mathrm{f}(v_a)$) are represented for one constant value of the screen-grid voltage V_{g2} and for a series of constant values of the control-grid voltage V_{g1} as well as a collection of working characteristics[4]). (Substantially the same applies for a tetrode.) In the case of the latter characteristics there was a loudspeaker connected to the anode circuit and a series of sinusoidal alternating voltages of increasing amplitude were applied to the control grid. In the absence of distortion the working characteristics would be pure ellipses.

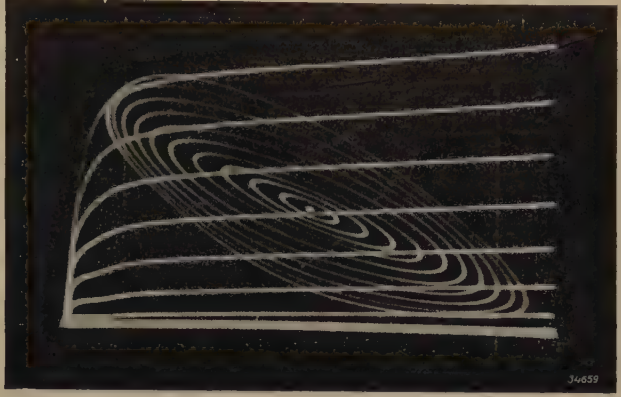

Fig. 2. Characteristics $i_a = \mathrm{f}(v_a)$ of a pentode. The elliptical curves are the paths described by the working point when the load impedance in the anode circuit consists of a loudspeaker and a sinusoidal alternating voltage with stepwise increasing amplitude is applied to the control grid. It is seen that at a certain value v_a there may be widely different values of i_a.

It is seen that with varying amplitude of the a.c. grid voltage a large area of the plane of the characteristics is covered and that at a certain value of the anode voltage various values of the anode current are possible. In order to avoid distortion — in so far as this arises through secondary emission — secondary emission has to be suppressed in the whole of this area. It will appear that this cannot very well be realized without applying other means at the same time.

Let us first suppose that the valve is of such a construction that with a high anode current there is a sufficiently strong space charge to counteract secondary emission; the trend of the potential v in the space between screen grid and anode then follows a curve similar to curve (*1*) in *fig. 3*. With equal potentials of screen grid and anode, but with a small anode current, the minimum however disappears completely (curve *2*, fig. 3). If it had been arranged for a sufficient minimum to be present already with small anode currents then it might well fall too low, even to about zero ("virtual cathode"), when large anode currents are flowing. In that event some of the electrons are no longer able to reach the anode and then again considerable distortion takes place.

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[4]) A method by which such characteristics can be recorded is described by A. J. Heins van der Ven, Testing amplifier output valves by means of the cathode-ray tube, Philips Techn. Rev. 5, 61-68, 1940.

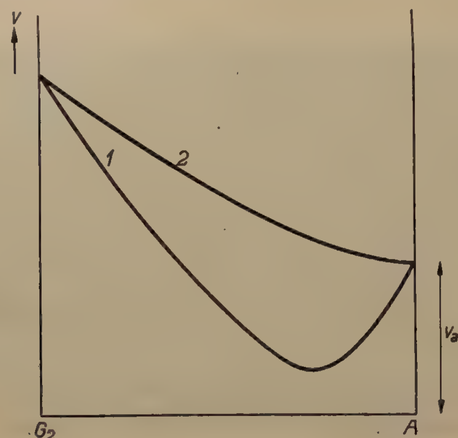

Fig. 3. The potential v between the screen grid (G_2) and the anode (A) of a tetrode with concentrated space charge. Curve *1* applies for a high anode current and curve *2* for a low anode current. In curve *2* the potential minimum has disappeared.

Such a state of affairs must be prevented at all costs. In the case of output tetrodes the usual practice is to mount screens between the screen grid and the anode which are connected to the cathode and to a certain extent counteract the transition of secondary electrons when the anode current is small. These screens are to be regarded as a simplified form of a suppressor grid, which we shall now consider.

Third grid

Secondary emission can also be suppressed by providing the tetrode with a third grid (making the valve a pentode) and connecting this so-called suppressor grid to a point where the potential is sufficiently lower than that of the anode [5].

The suppressor grid should preferably have a large pitch. Therefore we have to differentiate between the (constant) voltage V_{g3} of the suppressor grid wires and the potential existing in the meshes of the suppressor grid. The instantaneous value of this potential depends upon V_{g3}, the anode voltage and the screen-grid voltage, and it may to a good approximation be identified with the effective potential $v_{3\mathrm{eff}}$ that has to be given to the plane of the suppressor grid.

With an efficient construction it can be arranged that when $V_{g3} = 0$ the value of $v_{3\mathrm{eff}}$ is kept lower than v_a for greatly differing values of v_a (*fig. 4*, curves *1* and *2*) independently of the anode current. It is this independency of the anode current that constitutes the great advantage of the pentode compared with a tetrode with concentrated space charge.

When v_a drops to very low values then, under the influence of the screen-grid voltage, $v_{3\mathrm{eff}}$ may become higher than v_a (curve *3* in fig. 4) and thus draw the secondary electrons away from the anode instead of returning them to it.

When $V_{g3} = 0$ the potential $v_{3\mathrm{eff}}$ is given by

$$v_{3\mathrm{eff}} = \frac{D_1 v_{2\mathrm{eff}} + D_2 v_a}{1 + D_1 + D_2} \quad \ldots\ldots\ldots \quad (1)$$

in which $v_{2\mathrm{eff}}$ = effective potential in the plane of the screen grid (which in practice can be taken as being approximately equal to V_{g2}), D_1 = "Durchgriff" (reciprocal of amplification factor) of the anode in the suppressor grid. Fig. 4 has been plotted for $D_1 = D_2 = 0.5$.

For the limit case where $v_a = v_{3\mathrm{eff}}$, when there is no longer any potential minimum, it follows from eq. (1) that

$$v_a \approx \frac{D_1}{1 + D_1} V_{g2} \quad \ldots\ldots\ldots \quad (2)$$

To keep this limit value of v_a as low as possible one must therefore make D_1 small. To this end one might make the suppressor grid out of closely wound wire turns, but then this would obstruct too much the passage of the electrons to the anode, and as the result the screen-grid current would be greatly increased at the cost of the anode current. Therefore one preferably applies a large pitch suppressor grid placed as far away from the screen grid as possible in order to keep D_1 small [6]).

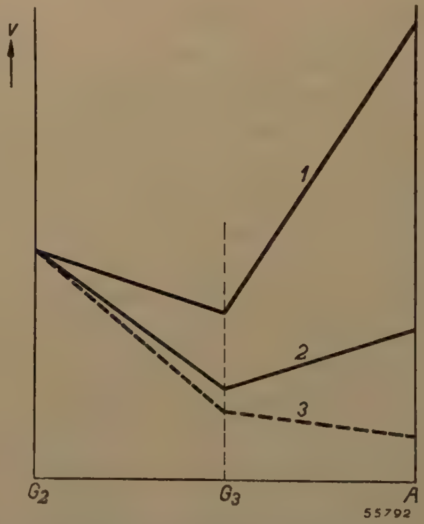

Fig. 4. The potential v between the screen-grid (G_2) and the anode (A) of a pentode in a plane passing midway between two wires of the suppressor grid. G_3 = suppressor grid at zero potential. Curve *1* applies for a high instantaneous value of v_a and curve *2* for a low ditto; in both cases there is a potential minimum in the meshes of G_3. This is no longer the case with curve *3*, applying for a still lower value of v_a.

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[5] See, e.g., British patent 287 958, priority 1926 (G. Holst and B. D. H. Tellegen.)

[6] For a more extensive treatise on these matters see J. L. H. Jonker, Pentode and tetrode valves, Wir. Engr. **16**, 274-286 and 344-349, 1939.

In order to suppress secondary emission at very low instantaneous values of anode voltage, at least at a high anode current, modern output pentodes use also a concentration of space charge between suppressor grid and anode, this being brought about by suitable choice of the shape, size and spacing of the electrodes. The suppression is then only inadequate in the area where low values of i_a and v_a occur simultanously, but, as will be seen from fig. 2, this zone lies outside the working range of the valve.

Combination of suppressor grid and prepared anode

We now come to the third means by which secondary emission can be counteracted, the coating of the anode with a layer which the electrons cannot easily leave. Here it may be a question either of the nature of the substance forming the layer or of the form of that substance, or the solution may be found in both. As regards its nature it is obvious to choose a substance having a coefficient δ of secondary emission lower than that of the anode, which usually consists of nickel plate (by δ is understood the number of secondary electrons leaving the bombarded electrode as an average per primary electron). The form of the substance is of still more importance than its nature: a flocculated form is highly favourable because the secondary electrons cannot easily emerge from the numerous cavities in and between the flakes ("labyrinth effect"). A distinction has to be made between the coefficient δ (just mentioned) relating to a flat plane of material in a non-porous state and the coefficient δ_{eff} referring to materials in some other form. With a nickel plate covered, for instance, with floccular soot, values of δ_{eff} can be reached 3 to 5 times as low as the δ value of pure nickel plate.

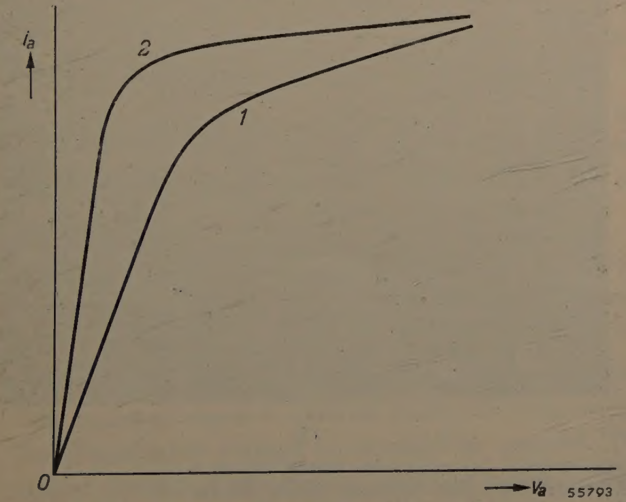


Fig. 5. Curve *1* is the characteristic $i_a = \mathrm{f}\,(v_a)$ of a normal pentode and curve *2* that of a pentode in which the anode has a coating with a low secondary emission.

When this remedy is applied in a well-designed pentode, the suppressor grid of which is of such dimensions that the characteristics (curve *1* in *fig. 5*) no longer show any irregularities that are to be attributed to secondary emission, then a prepared anode as mentioned above does nevertheless give an improvement (curve *2* in fig. 5): the flat part of the characteristic is more horizontal and the bend in the curve is shifted to higher values of i_a and lower values of v_a. This means that for the working characteristic larger ellipses can be allowed — thus the valve can yield a greater output — before a certain amount of distortion arises.

To explain this rather unexpected effect we have to consider more closely the properties of secondary electrons.

Velocity distribution of secondary electrons

When a conductor is bombarded with electrons having a velocity V_1 electron volts then the emerging electrons have various velocities V between 0 and V_1. *Fig. 6*, relating to a nickel plate, shows the curve of the velocity distribution of these electrons in the case of primary electron velocities V_1 of 50, 150 and 250 eV. In every case a large proportion of emerging electrons are found to have a velocity of about 5 eV. These are the "real" secondary electrons. The curves show a peak at a velocity but little lower than V_1; this peak corresponds to the reflected electrons, i.e. the primary projectiles which have almost entirely retained their energy after the collision. (As already remarked in the introduction, the reflected electrons, too, are usually reckoned to belong to the secondary electrons.) The part of the curve lying between the two peaks represents those electrons which through the collision have lost a larger or smaller part of their energy.

From fig. 6 it appears that with lower velocities of the primary electrons the number of "real" secondary electrons relatively decreases whilst the number of reflected electrons increases.

In a tetrode or pentode it is particularly these reflected electrons which, owing to their high velocity, are easily capable of overcoming the repellent action of the potential drop between anode and screen grid and which are thus absorbed by this grid. This holds for any value of v_a, so that, as opposed to the "real" secondary electrons, the reflected electrons cannot give rise to any inflection points in the characteristics $i_a = \mathrm{f}(v_a)$. What does happen, however, according to fig. 6, is that with falling anode voltage the number of reflected electrons becomes greater, which means that as v_a decreases the characteristic assumes a steeper slope, as illustrated by curve *1* in fig. 5. When v_a drops to such low values to be about equal to or less than $v_{3\mathrm{eff}}$ then the real secondary electrons also pass over to the screen grid and thus bring about a further reduction of i_a. In this way bends may arise in the characteristic which cause distortion when they lie within the working range.

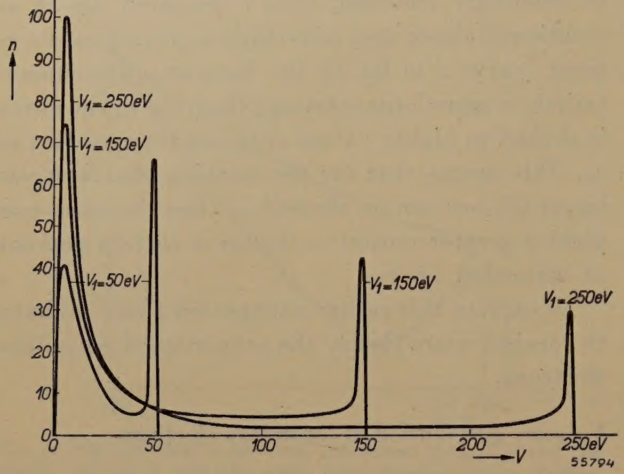


Fig. 6. Velocity distribution of secondary electrons obtained when a nickel plate is bombarded with primary electrons having a velocity V_1 of 250, 150 and 50 eV respectively. As function of the velocity V a quantity n is plotted which is proportional to the number of secondary electrons having the velocity V; $n = 100$ has been taken arbitrarily for the highest peak of the three curves. The peaks at $V \approx 5$ eV correspond to the "real" secondary electrons, the peaks at $V \approx V_1$ correspond to reflected electrons.

For the case of nickel coated with a porous layer of soot we have the curves of *fig. 7*. Here it is seen that the fast electrons have been relatively greatly reduced in number. The real secondary emission, too, is reduced: for the pure nickel plate $\delta = 1.2$ and for the plate covered with soot δ_{eff} is only 0.36 (in both cases at $V_1 = 250$ V). When such a coating is applied to the anode of a pentode the resultant reduction of the secondary emission manifests itself in the previously mentioned improved shape of the characteristic $i_a = \mathrm{f}(v_a)$ (curve *2* instead of curve *1*, fig. 5).

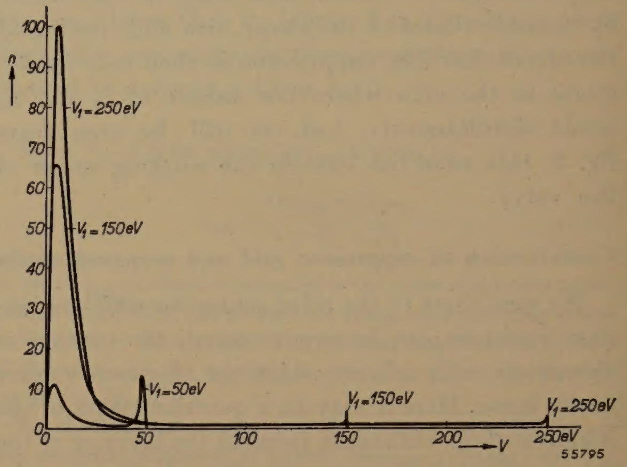
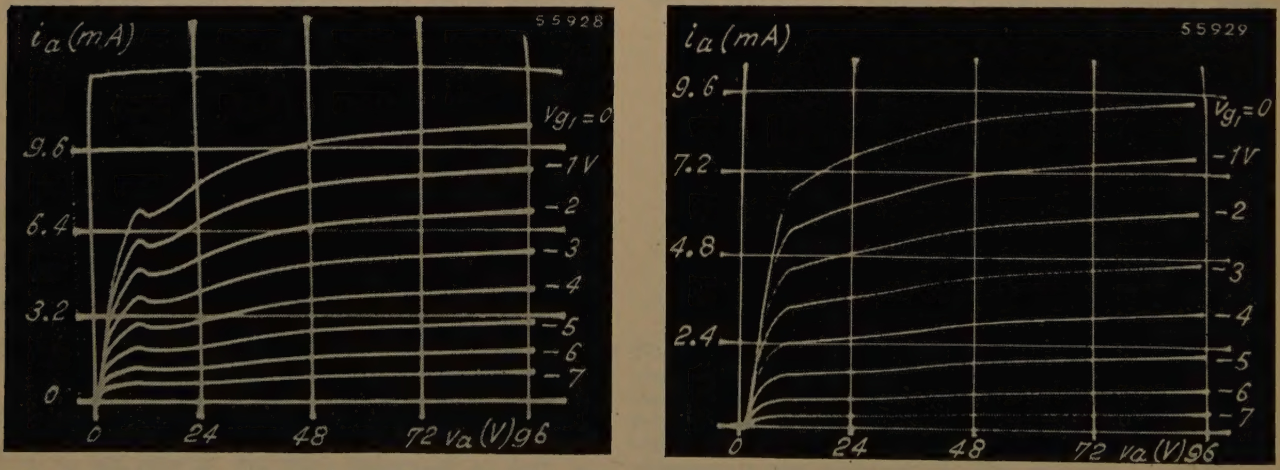


Fig. 7. As in fig. 6, but for the case where the nickel is coated with a porous layer of soot. The reflected electrons have lost much importance. To compare the ordinates with those of fig. 6 it is to be borne in mind that the area between a curve and the V-axis is a measure for the (effective) coefficient of secondary emission. For $V_1 = 250$ eV in fig. *6* $\delta = 1.2$ and in this figure $\delta_{\mathrm{eff}} = 0.36$, from which it follows that the "real" secondary emission has likewise been reduced.

Output pentode for battery supply

The difficulties arising from secondary emission are felt most strongly in the case of valves which are supplied from an anode battery. The supply voltage is usually rather low (45 to 135 V) and, as we have seen from fig. 6, the fast reflected electrons are most



a *b*

Fig. 8. Charcteristics $i_a = \mathrm{f}(v_a)$ of two similar pentodes, (*a*) with bare anode, (*b*) with coated anode, at $V_{g2} = 90$ V and $V_{g3} = 0$. In (*a*) decided secondary emission occurs, whereas in (*b*) it does not. Attention is drawn to the very low value of v_a at which in the case (*b*) the bend occurs in the characteristics. The latter series of characteristics apply for the new output pentode (DL 41) designed for battery feed.

numerous when the voltage is low. Moreover, having regard to the useful life of the battery, a low mean anode current has to be used, so that the contributory help from the action of the space charge is for the greater part lost. It is therefore desirable, in particular for output valves fed from an anode battery, to reduce the number of secondary electrons and especially the reflected ones. This is best effected by coating the anode with porous soot or a similar material as referred to above.

The result is to be seen in *fig. 8*. In fig. 8*a* the characteristics $i_a = f(v_a)$ of a pentode with a normal nickel anode are shown [7], and in fig. 8*b* we have the characteristics of a similar pentode provided with a prepared anode (type DL 41). Particularly at low anode voltages the latter characteristics have a very much more favourable shape. This is manifest in the maximum output that the DL 41 valve can yield with a given degree of distortion: with 10% distortion it is about 260 mW compared with 200 mW from a similar valve without anode coating [8].

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[7] Recorded with a cathode-ray tube, cf. footnote [4], but with an improved apparatus by means of which also a network of coordinates is plotted automatically.

[8] Measured with a battery voltage of 120 V, $V_{g2} = 120$ V and a load resistance of 24000 Ω in the anode circuit.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS OF THE N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk can be obtained free of charge upon application to the Administration of the Research Laboratory, Kastanjelaan, Eindhoven, Netherlands.

R 78: W. J. Oosterkamp: The heat dissipation in the anode of an X-ray tube, II. Loads of short duration applied to rotating anodes (Philips Res. Rep. 3, 161-173, 1948, No. 3).

The maximum temperatures occurring in the target of an X-ray tube, or in the copper backing behind it, are computed on the same principles as adopted in the first part of the paper (see these abstracts, No. **R** 71), but this time for a rotating anode. It is shown that for the short exposures the lateral dissipation of heat may be disregarded. Exposures lasting less than one revolution and those lasting several revolutions are dealt with separately. From the resulting formulae the rating capacity is computed as a function of the time of exposure and is found to agree closely with life-tests.

R 79: Balth. van der Pol: Mathematics and radio problems (Philips Res. Rep. 3, 174-190, 1948, No. 3). The contents of this paper are the same as that of No. **1759.**

R 80: A. van Weel: Developments in radio-receiver circuits for the ultra-short-wave range (Philips Res. Rep. 3, 191-212, 1948, No. 3).

Push-pull converter stages are described in which the antenna signal is used in push-pull whereas the local-oscillator voltage is supplied in parallel. The input circuit of the mixing valve is tuned to both frequencies at the same time. Special measures to prevent the local-oscillator voltage from penetrating into the antenna are described. Diode and triode mixing stages are dealt with in detail. It proves to be possible to design self-oscillating triode converter stages, as a consequence of which a separate local-oscillator valve is made superfluous and the noise that might be induced by this valve is eliminated. (Partly dealt with in Philips Techn. Rev. **8**, 193-198, 1946.)

R 81: C. J. Bouwkamp: On the theory of coupled antennae (Philips Res. Rep. 3, 213-226, 1948, No. 3).

The purpose of this paper is to provide formulae for the self and mutual impedances of an antenna system consisting of two identical, centre-fed, parallel, cylindrical wires. Tables of auxiliary functions are given, facilitating numerical investigation in the range $1 < 2\pi l/\lambda < 2$, $0 < d/l < 2$, $\Omega = 2\ln(2l/a) > 10$, where $2d$ is the length of each antenna, d their mutual distance, $2a$ the wire diameter, and λ the wavelength.

R 82: W. W. Boelens: Valve characteristic giving linear modulation when a feedback resistor is inserted in the cathode lead (Philips Res. Rep. 3, 227-234, 1948, No. 3).

It is well known that a high-frequency voltage can be modulated with a low-frequency signal by applying both voltages to the control-grid of a radio valve. If the I_a-V_g characteristic is quadratic there will be no distortion in the modulation. The available valves, however, generally have a characteristic that is not quadratic. A feed-back resistance in the cathode lead was found to improve the linearity of the modulation; there was an optimum value for this resistance.

The theoretical question was put as to what valve characteristic would give linear modulation with a given feed-back resistance. This problem is solved in the present article and a simple answer is found. Two graphs are drawn for the characteristics with the feed-back resistance as a parameter.

R 83: J. M. Stevels: Some investigations in the system Na_2O-PbO-TiO_2-SiO_2 (Philips Res. Rep. **3**, 235-238, 1948, No. 3).

A number of diagrams are given showing the vitreous and the crystalline regions in the systems Na_2O-SiO_2-Pb_2TiO_4, Na_2O-SiO_2-$PbTiO_3$ and Na_2O-SiO_2-$PbTi_2O_5$.

R 84: F. L. H. M. Stumpers: Noise in a pulse-frequency-modulation system (Philips Res. Rep. **3**, 241-254, 1948, No. 4).

The optimum filter for pulse-frequency-modulation is derived for any given pulse form and large signal-to-noise ratio. For some special pulse forms she signal-to-noise ratio is calculated and it is shown that normal frequency modulation gives a better result. A method is given for the calculation of the noise spectrum, which is valid for all signal-to-noise ratios, though the intricacy of the formula restricts the possibilities for application. The noise threshold is estimated and the suppression of the modulation by noise is calculated.

R 85: A. van der Ziel and A. Versnel: The noise factor of grounded-grid valves (Philips Res. Rep. **3**, 255-270, 1948, No. 4).

Measurements are given of the noise factor of grounded-grid valves at 7.25 m wavelength. In such valves part of the output noise currents also flows through the input circuit, thus giving rise to a partial noise suppression.

A general theoretical treatment of the problem is given, dealing with the influence of circuit losses, dielectric losses, transit-time effects, field inhomogeneities, and of an independent noise current in the output circuit which is not present in the input circuit. This theory is applied to grounded-grid triodes and to grounded-grid pentodes and secondary-emission valves. The experimental results for grounded-grid triodes agree qualitatively with theory. The experiments prove that a loose antenna coupling is favourable for a low noise factor, but that complete suppression of noise is impossible.

R 86: J. D. Fast: The equilibrium between liquid silica and liquid iron (Philips Res. Rep. **3**, 271-280, 1948, No. 4).

The equilibrium between liquid SiO_2 and liquid Fe is computed with the aid of experimental data from the literature and thermodynamic considerations. The possible sources of error are discussed.

R 87: J. L. Meijering: Retrograde solubility curves, especially in alloy solid solutions (Philips Res. Rep. **3**, 281-302, 1948, No. 4).

Using Gibbs's entropy of mixing and Richards's rule for the entropy of fusion of metals, a graph is derived thermodynamically, which serves to predict whether a solidus (curve representing the concentrations of mixed crystals in coexistence with liquid) is retrograde, i.e. shows a maximum with respect to the temperature axis. The result for the metals is checked with experimental data and is extended to other retrograde phenomena (solidus curves in systems of non-metals, retrograde allotropic transformations and retrograde solubility in liquids).

R 88: W. J. Oosterkamp: The heat dissipation in the anode of an X-ray tube, III. Continuous loads (Philips Res. Rep. **3**, 303-317, 1948, No. 4).

In this paper (cf. **R 78** and **R 71**) the effect of continuous loads is considered in the case of both a stationary and a rotating anode.